WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JUNE 1956

VOL. 33 No. 6 · THREE SHILLINGS AND SIXPENCE

Photograph by courtesy of British Insulated Callenders' Cables, Ltd.

Avo in action..

The checking of a busbar installation in a large steel works is an interesting example of the important contribution of the world-famous AvoMeter to the maintenance of electrical power plant. AvoMeters will be found wherever electricity is used in industry.

Model 40 *Universal* AvoMeter

A self-contained multi-range A.C./D.C. instrument providing 40 ranges of direct readings of current, voltage and resistance on a 5-inch hand calibrated scale fitted with an anti-parallax mirror. Range selection is by two rotary switches, for A.C. and D.C. respectively, and a press button provides additional ranges by halving the current and voltage ranges shown on the switch knobs. The total resistance of the meter is 200,000 ohms.

The instrument is compact and portable, and is protected by an automatic cut-out against damage through the inadvertent application of overload.

Power Factor and reactive kV.A. can be measured in A.C. circuits by means of an external accessory (the Universal AvoMeter Power Factor & Wattage Unit).

Size: 8 × 7¼ × 4½ ins.
Weight: 6¼ lbs.

List Price
£19 : 10s.

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Sole Proprietors and Manufacturers:—

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A40/3

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This is our Model ASR-1150 Automatic Voltage Step Regulator. It ideally fills the need for a cheap, small and light Stabiliser. Although it measures only 8½ in. × 4½ in. × 5 in., weighs as little as 11 lb., and costs only £24 net, it has a performance fully equal to any similarly rated Automatic Stabiliser of the resonated, saturated core type, without any of the disadvantages.

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ENGLISH ELECTRIC VALVE CO. LTD. *Waterhouse Lane, Chelmsford*
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AP 300/40

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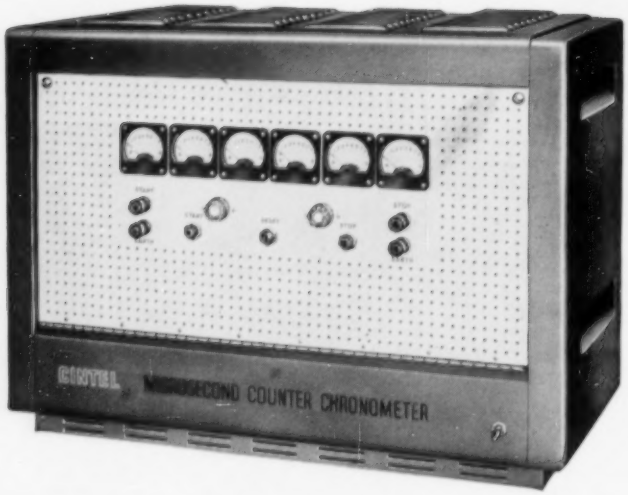

This equipment is designed to measure small intervals of time to a high order of accuracy and two ranges are provided:

(1) 1 μsec. to 1 sec. in steps of 1 μsec.

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Accuracy of each range is better than ± 0.005% ± the step interval.

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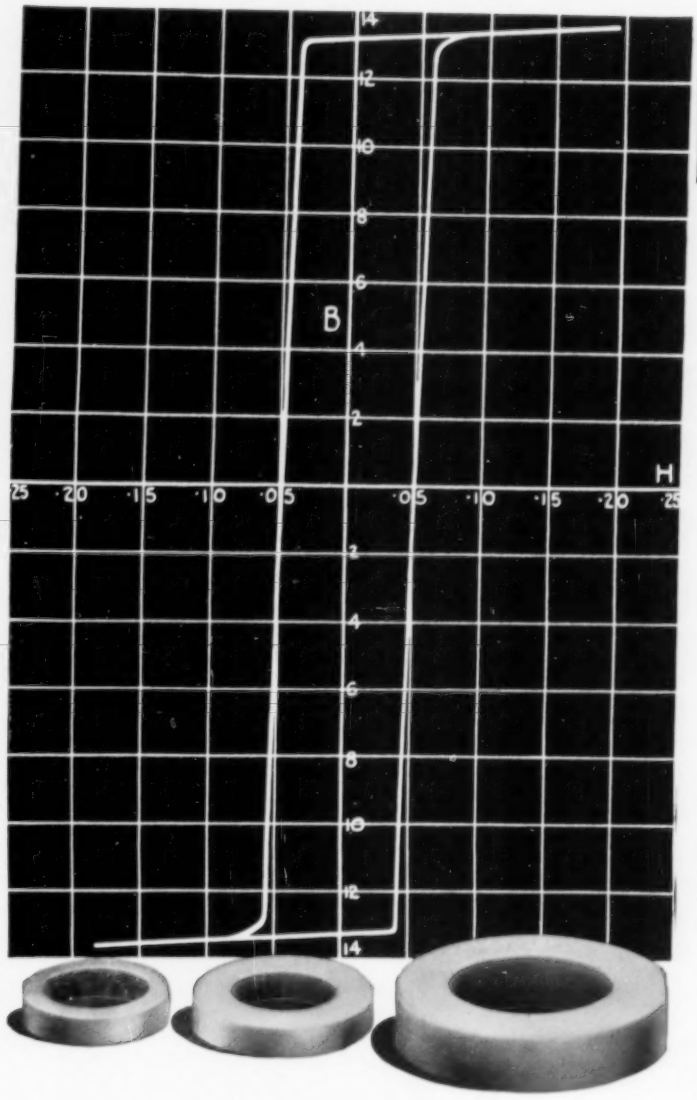

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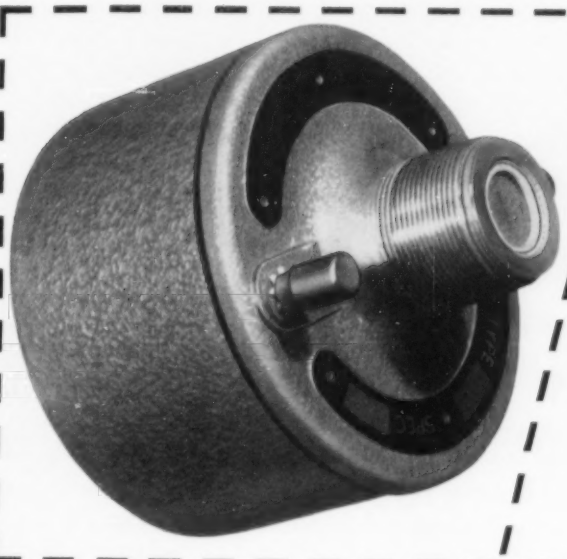

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PRESSURE UNIT

SPECIFICATION

Power Handling Capacity	10 watts peak
Voice Coil Impedance	15 ohms
Frequency Response	120–9000 c.p.s.
Flux Density	12,000 gauss
Pole Piece	1·5″ diameter

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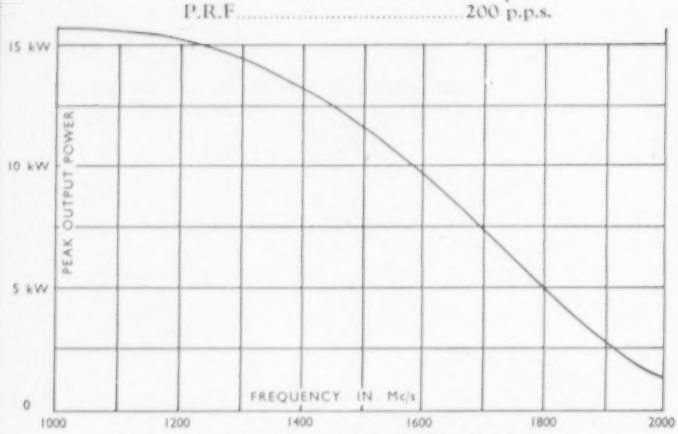

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'Welmet' Metal Film Resistors
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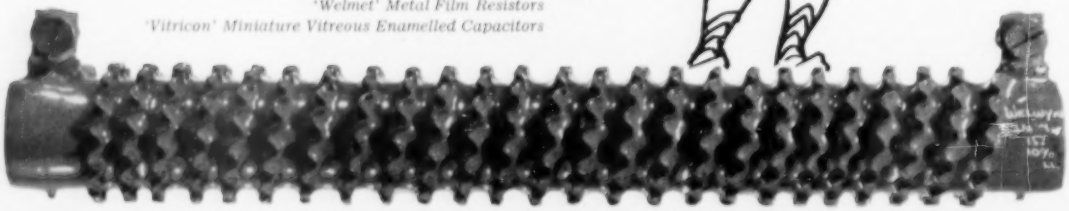

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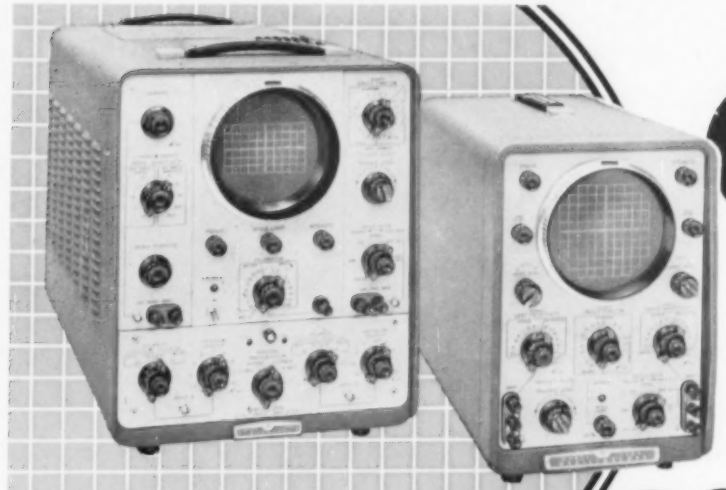

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2

all new

OSCILLOSCOPES

-hp- 130A Low Frequency Oscilloscope

High sensitivity, dc to 300 kc. Sweeps 1 μsec/cm to 15 sec/cm.

-hp- 150A High Frequency Oscilloscope

Dc to 10 Mc. Plug-in preamplifiers. Sweeps 0·02 μsec/cm to 15 sec/cm.

As a result of a totally new design philosophy, -hp- 130A and 150A Oscilloscopes set revolutionary standards for oscilloscope usefulness, convenience and rugged dependability. The instruments' wide versatility is indicated in the specifications at left. Their greater convenience and reliability is inherent in such unique features as: Universal automatic triggering system wherein one preset adjustment provides optimum triggering for almost all conditions. Unitized circuits, easily isolated for testing. Etched circuits, mounted on translucent plastic for "see-through" serviceability. Complete accessibility of all circuits and tubes. Highest quality components. Ultra-conservative design, with circuits operating well below rating. Concentric, colour-coded controls, grouped by function and simplified. Direct sweep time selection—no mental gymnastics. A new low capacitance clip-on probe (for -hp- 150A) has 10 megohm impedance.

Get complete data on the most important oscilloscope improvements in history! Call your -hp- representative now, or write direct!

BRIEF SPECIFICATIONS

-hp- 130A

Sweep Range: 1 μsec/cm to 15 sec/cm.

Calibration: 21 sweeps: 1-2-5-10 sequence, 1 μsec/cm to 5 sec/cm. 5% accuracy.

Triggering: Internal, line voltage or external 2 v or more. Pos. or neg. slope, + 30 to − 30 v trigger range.

Preset Trigger: Optimum setting for automatic stable triggering.

Input Amplifiers: (Similar Vert. and Horiz. Amps.) Sensitivity 1 mv/cm to 50 v/cm; 14 ranges, continuous vernier. Pass band dc to 300 kc.

Amplitude Calibration: 1 kc square wave. 5% accuracy.

Price: $450.00.

-hp- 150A

Sweep Range: 0·02 μsec/cm to 15 sec/cm.

Calibration: 24 sweeps: 1-2-5-10 sequence, 0·1 μsec/cm to 5 sec/cm. 3% accuracy.

Triggering: Internal, line voltage or external 0·5 v or more. Pos. or neg. slope, + 30 to − 30 v trigger range.

Preset Trigger: Same as -hp- 130A.

Horizontal Amplifier: Magnification 5, 10, 50, 100 times. Vernier selects any 10 cm part of sweep. Pass band dc to over 500 kc. Sensitivity 200 mv/cm to 25 v/cm.

Vertical Amplifier: Pass band dc to 10 Mc. Optimum transient response and rise time less than 0·035 μsec. Signal delay of 0·25 μsec permits leading edge of triggering signal to be viewed.

Amplitude Calibration: 18 calib. voltages, 2-5-10 sequence, 0·2 mv to 100 v peak-to-peak. Accuracy 3%. Approx. 1 kc square wave, rise and decay approx. 1·0 μsec.

Prices: -hp- 150A High Frequency Oscilloscope, $1,000.00.
-hp- 151A High Gain Amplifier, $100.00.
-hp- 152A Dual Channel Amplifier, $200.00.

Data subject to change without notice.
Prices f.o.b. factory.

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Represented by

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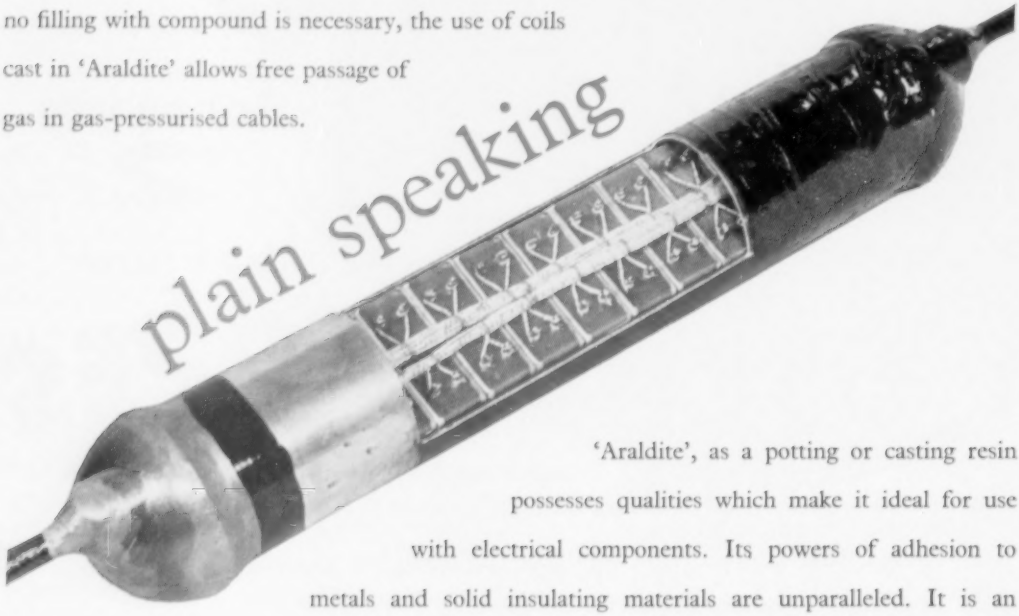
These Mullard Loading Coils represent the latest technique in splice loading of voice frequency cables, whereby the exact number of coils may be installed to meet current requirements. The coils are cast in 'Araldite' epoxy resin which gives complete protection against climatic conditions and permits the use of shapes which will occupy the minimum of space in the joint. As no filling with compound is necessary, the use of coils cast in 'Araldite' allows free passage of gas in gas-pressurised cables.

plain speaking



'Araldite', as a potting or casting resin possesses qualities which make it ideal for use with electrical components. Its powers of adhesion to metals and solid insulating materials are unparalleled. It is an excellent insulator, tough, flexible, highly resistant to 'tracking' and has a low power factor. It does not cause corrosion, provides complete protection against tropical conditions and complies with service requirements relating to sealing of components.

'Araldite' epoxy resins have a remarkable range of characteristics and uses.

They are used
- ★ **for bonding metals, porcelain, glass, etc.**
- ★ **for casting high grade solid insulation.**
- ★ **for impregnating, potting or sealing electrical windings and components.**
- ★ **for producing glass fibre laminates.**
- ★ **for producing patterns, models, jigs and tools.**
- ★ **as fillers for sheet metal work.**
- ★ **as protective coatings for metal, wood and ceramic surfaces.**

Please write now for full descriptive literature on Araldite Epoxy Resins.

'Araldite' **epoxy resins**

'Araldite' is a registered trade name.

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for precise control of superior electronic equipment

—moulded track potentiometers by **Plessey**

COMPACT

The physical confines and electrical parameters of modern electronic equipment makes unprecedented demands on the component manufacturer. His product must almost invariably be smaller than its predecessor and more efficient and durable. These potentiometers provide an outstanding example of what can be achieved to meet the present requirement.

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Moulded track potentiometers by Plessey possess a high standard of stability within a wide range of operational temperatures and can be stored for extended periods without deterioration. They can be manufactured to conform, within strictly specified limits, to a designed pattern of values at various positions on the track.

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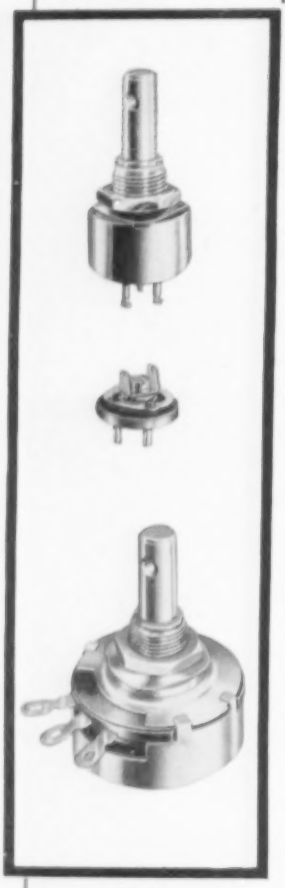
The outstanding reliability of these potentiometers is a tribute to the Plessey Research and Development engineers whose efforts to meet the stringent Services requirements and the critical demands of the Commercial market have resulted in products of superior design and function.



Miniature Type M H 2

A high quality potentiometer for commercial laboratory and Services application. Services Specification RCS122. The standard version is fitted with spindle and panel humidity seals.

GENERAL SPECIFICATION

Rating: ambient 70°C	¼ watt
Temperature Cat.	40/70°C
Humidity Class	H2
Minimum Resistance	50 ohms max.
Overall Diameter	23/32 in.
Overall depth of control from mounting face, including tags	⅜ in.

Sub Miniature Type G

Type Approved to Certificate No. 974 Issue 1, Ref. TA1373, this pre-set, open-type control is of particular interest to industrial communication and Service equipment manufacturers.

GENERAL SPECIFICATION

Range Linear ±20%	1,000 ohms to 2 megohms
Rating: ambient 70°C	¼ watt total track rating
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Spindle and Panel Sealed Type E H 2

For military and para-military applications, this control is particularly resistant to conditions of extreme humidity.
Type Approved to Certificate No. 906, Issue 2, Ref. TA1230.

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Humidity Class	H2
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Overall Diameter	1-5/32 in.
Overall depth of control from mounting face	21/32 in.

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ALUMINISED C.R. TUBES

for tube replacements

The G.E.C. 14 in. Cathode Ray Tube type 7203A has the following attractive features:—

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2) A narrow beam angle *triode* gun giving:—

(a) Uniform focus over the entire screen area.

(b) Small spot size.

(c) Freedom from astigmatism.

3) A high reflectance aluminium backing to the fluorescent screen giving:—

(a) 70% increase in picture brightness.

(b) High contrast daylight viewing.

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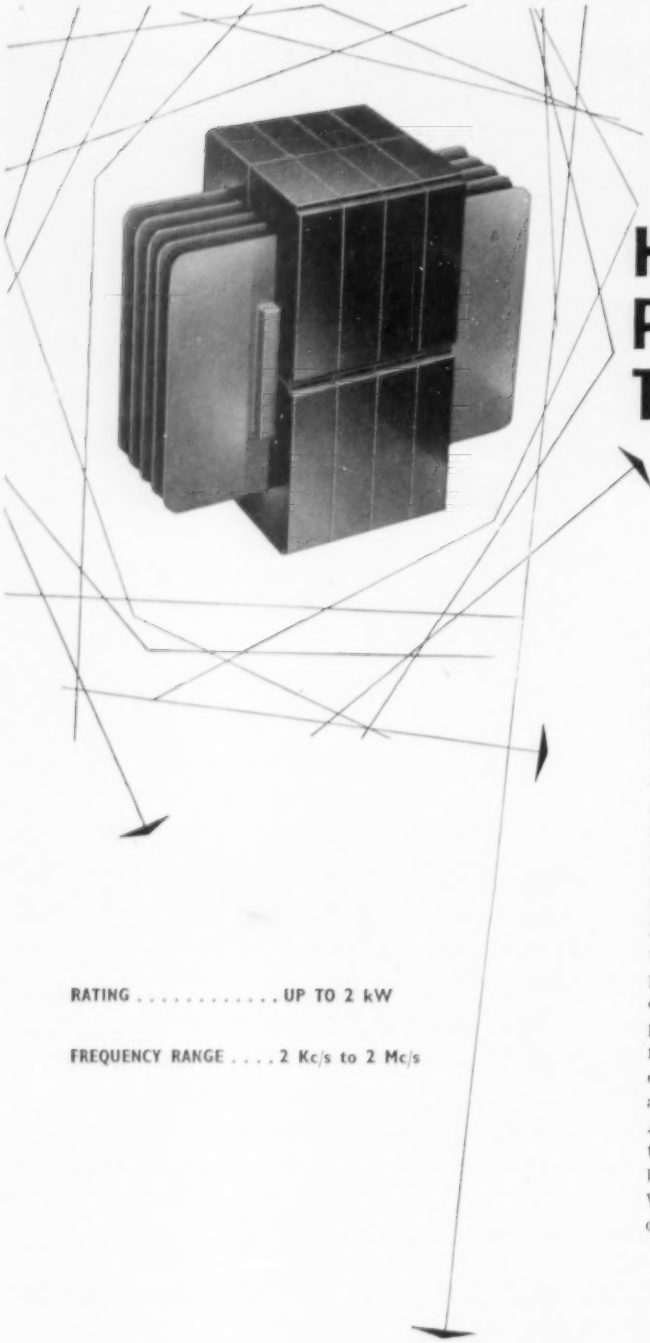

H.F. POWER TRANSFORMERS

H.F. power transformers of outstanding efficiency are the latest additions to the Mullard range of high quality components designed around Ferroxcube magnetic cores.

Utilising the unique characteristics of Ferroxcube to the full, Mullard H.F. transformers are smaller, lighter, and less costly than transformers using alternative core materials. These advantages are particularly marked in transformers required to handle powers of up to 2kW, between the frequency range 2kc/s to 2Mc/s.

Mullard transformers are already finding wide use in applications as diverse as ultrasonic H.F. power generators and aircraft power packs operating from an aircraft's normal A.C. supply. In the latter application, the low leakage field of Ferroxcube can eliminate the need for external screening, thereby reducing the size and weight of the transformer even further.

As with all Mullard high quality components, these H.F. power transformers are designed and built to engineers' individual specifications. Write now for details of the complete range of components available under this service.

RATING . . . . . **UP TO 2 kW**

FREQUENCY RANGE . . . . **2 Kc/s to 2 Mc/s**

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The Journal of Radio Research and Progress

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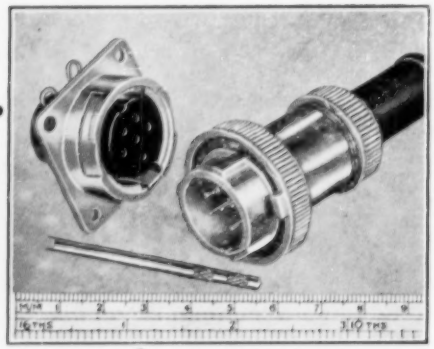
Multipole Screened Plugs and Sockets



L766/FP Free Plug
L766/CS Fixed Socket

The "Belling-Lee" range of screened plugs and sockets is designed for use with braided and insulated multicore cables. The gold-plated plug and socket assemblies are interchangeable in their housings, the latter being the same as those used for our L.764 series of coaxial plugs and sockets except that the insulant is nylon filled phenolic material. The protective cover is also common to all types.

The exploded view shows the various piece parts of the free member

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3-pole	*L.1326/FP	L.1326/CS
4-pole	L.765/FP	L.765/CS
6-pole	L.766/FP	L.766/CS
12-pole	L.767/FP	L.767/CS

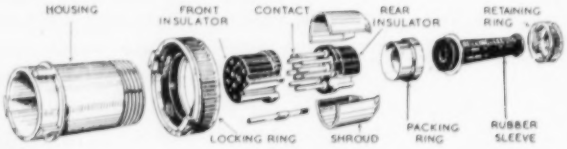
* Waterproof with black anodised finish, to Lloyd's specification of requirements for shipping. U.K. Pat. 715574.

Protective cover L.764/CA/ anodised or polished L.764/C/Polished.

L1327/FP Free Plug
L1327/CS Fixed Socket

Contact Assembly	Max. Cable dia.
2-pole 15 amp.	0.38 in.
3-pole 15 amp.	0.38 in.
4-pole 19 amp.	0.45 in.
6-pole 5 amp.	0.32 in.
12-pole 5 amp.	0.41 in.

L765/FP Free Plug
L765/CS Fixed Socket



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Physical Society's Exhibition

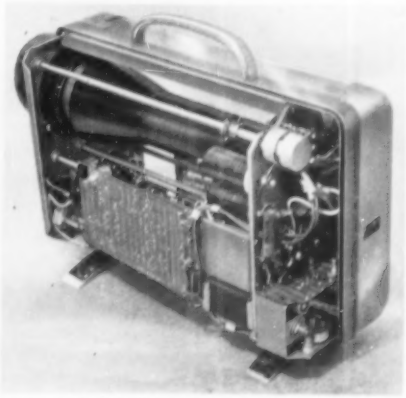
THIS 40th exhibition of scientific instruments and apparatus, which was held from 14th to 17th May, has so grown that, this year, it occupied both the Old and the New Horticultural Halls. In spite of this, it was at times uncomfortably crowded but it was, nevertheless, a great improvement on last year, when only the New Hall was used. We remarked last year that the break with the tradition of holding it at Imperial College had made the exhibition lose much of its distinctive character. Oddly enough, and most pleasantly, it seemed this year to regain much of its old character and we no longer felt strangers in a strange exhibition.

One of the biggest and most crowded of the exhibits was the one arranged by the Colour Group of the Physical Society to demonstrate some of the principles of colour television. It included 20 main items starting with trichromatic vision, synthesis of colours by three-colour additive mixture and the specification of colour. The additive mixture of phosphor colours and of colour images were demonstrated by an analogy drawn between the eye and colour television. Colour analysis and synthesis, gamma correction and dichroic mirrors were explained, while a projection system, the R.C.A. tricolour tube, and the Lawrence tube were demonstrated. Field, line and dot-sequential scanning systems were also demonstrated as well as the separation of chrominance and luminance, colour discrimination in small fields and chrominance coding. The final exhibits covered compatibility, tolerance to colour balance and alternative luminance-chrominance transmission systems. It should, perhaps, be made clear that the demonstrations did not take the form of true pictures, but of artificial patterns.

Many of the main items included several sub-items and the exhibit as a whole was one of those outstanding ones which deserved a longer life than the four days it was given. Most people with any interest in colour television could have spent some profitable hours studying it closely, but too many people wanted to do so to enable anyone to derive full benefit from it.

Mullard showed several applications of a transistor operating under 'avalanche' conditions. It then functions in a manner which can be likened, perhaps rather crudely, to a thyratron. The collector is operated at a high voltage so that holes flowing to the collector reach a high enough velocity to ionize atoms near the collector. The electrons produced by this process flow in the base circuit and these produce a bias tending to increase the action.

The device can be used to produce a sudden



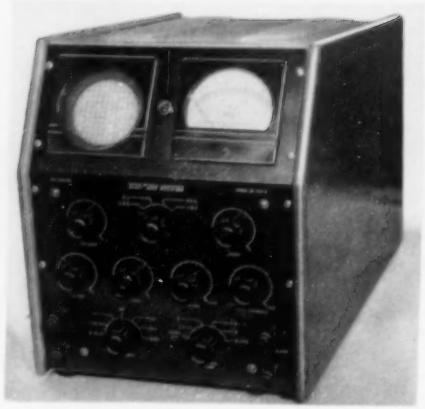
Mullard transistor oscilloscope.

discharge of a capacitor. With a 70-V supply, the transistor can produce a current pulse of 100-mA amplitude and 0·05-μsec duration. The voltage is of 60-V amplitude and of more or less sawtooth waveform. A repetition frequency of up to 1 Mc/s is possible.

A use which the transistor may fill in the future was also shown by Mullard in an experimental oscilloscope using only transistors in the deflection and power-supply circuits. Operating from a 12-V battery, the consumption is about 6 W and the h.t. supplies of 65 V and 1·2 kV are provided by ringing-choke circuits. The Y amplifier has a gain of 32 dB and a response from 30 c/s to 50 kc/s. It is, of course, a simple instrument and in this is quite unlike the majority of modern oscilloscopes which seem to be getting more and more complicated. To a large extent this is due to the trend towards making them precision measuring instruments rather than merely waveform indicators, as so many early types were.

Solartron type CD 523 "Solarscope".

An example of a modern general-purpose instrument is the Cossor 1058. It has a 4-in. tube with post-deflection acceleration and the Y amplifier has a response extending from zero to 6 Mc/s with a gain sufficient to provide a deflection sensitivity of 0·25 V/cm. The timebase is either free-running or triggered and the X amplifier provides for a sweep expansion of up to five times. Both voltages and time calibration facilities are included.



British Physical Laboratories Voltascope RS 900.

The Solartron CD 523 is another general-purpose type with a bandwidth of zero up to 10 Mc/s at a sensitivity of 10 V/cm. The amplifiers have their gain stabilized by feedback and the timebase is of the modified Miller type with a velocity range of 10 cm/μsec to 1 cm/sec in seven steps.

A smaller instrument is the combined oscilloscope and valve voltmeter of British Physical Laboratories. The valve voltmeter covers d.c. and a.c. up to 500 V, while the oscilloscope has a $2\frac{3}{4}$-in. tube and a gas-triode form of timebase. Both are intended mainly for use at frequencies up to about 100 kc/s.

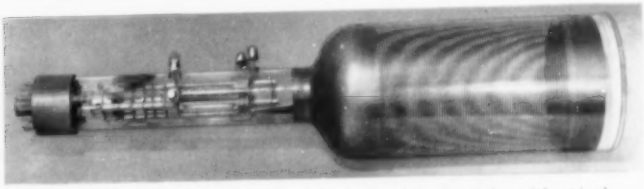
A much more specialized instrument is the Mullard L.140 which is designed for the measurement of pulse waveforms. A cathode-follower

Furzehill oscilloscope type O 130.

input probe is used and it can be biased off so that only the upper parts of a pulse waveform are passed, thus facilitating the accurate measurement of small *changes* of pulse amplitude. It is claimed that the cumulative error of time-display measurements does not exceed 3% for time intervals of 2 μsec to 150 msec.

An example of a very highly-specialized oscilloscope is the Cintel model, which is suitable for single-shot transient work. It has a tube with four post-deflection accelerators operating at up to 24 kV. There are no amplifiers and the signal is applied to the Y plates via a 72-Ω cable, the frequency response being maintained to 500 Mc/s, and the sensitivity is 50 V/in.

Among apparatus for use with the oscilloscope, the beam switch showed by Furzehill Laboratories is of interest. In effect, it converts a single-beam oscilloscope into a double-beam. It comprises a pair of amplifiers with cathode-follower input stages and a switching-voltage generator which may be synchronized by either input. The amplifiers cover 10 c/s to 5 Mc/s.



20th Century Electronics post-deflection accelerator tube with spiral accelerator.

A noteworthy development in oscilloscopes is the increasing use of c.r. tubes with one or more post-deflection accelerators in order to secure a bright trace while retaining high deflection sensitivity. A prototype tube of this kind was shown by 20th Century Electronics. It is unusual in that the accelerator takes the form of a spiral high-resistance coating on the wall of the tube. Some 8–10 kV is applied across its ends with the result that the accelerating field increases uniformly with distance along the tube.

Signal sources are always important in test and measurement. Probably stimulated by the introduction of f.m. broadcasting on Band II, quite a few frequency-modulated signal generators were shown.

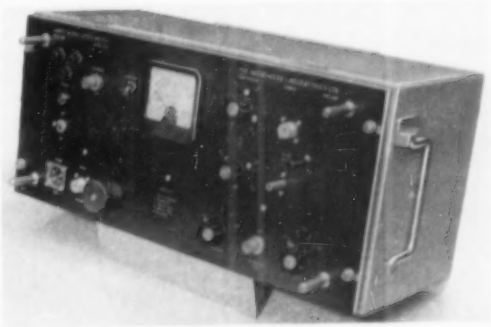
The Advance type 60 a.m./f.m. signal generator covers 4–225 Mc/s with a maximum output of 100 mV in 75 Ω. An internal source permits either a.m. or f.m. to be used and, by the addition of an external source, the carrier can be simultaneously modulated in both ways. The r.f. oscillator operates on fundamental frequencies and is followed by a tuned amplifier; the attenuator has a range of 100 dB.

Nagard delayed square-pulse generator, type 5001.

The Airmec signal generator type 204 covers 2–320 Mc/s. Internal modulation at 1 kc/s is provided and can be either a.m. or f.m. Whichever is used, the other can be obtained simultaneously by employing an external source. The deviation on f.m. is up to 80 kc/s on the lowest frequency range and up to 320 kc/s on the highest. A crystal calibrator is included and the output is up to 0·2 V.

The Hatfield Instruments LE 250 B signal generator is intended for use with an oscilloscope display of the receiver output. It is designed for work on f.m. receivers in the 81–105 Mc/s range, but there is a coverage of 3–20 Mc/s for i.f. alignment. The carrier can be frequency-modulated by a 400-c/s sine wave with a deviation of up to 75 kc/s or a 25-c/s sawtooth with a deviation of 0·5 Mc/s or 1 Mc/s.

In other fields other waveforms are needed, and an example of quite a different form of instrument is the Nagard wide-range delayed

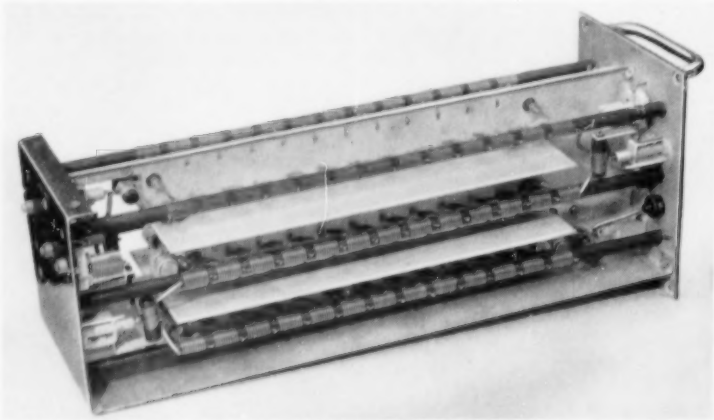


Wayne Kerr video noise-level meter, type M 131.

square-pulse generator. The main output is a pulse or square wave with a controllable amplitude calibrated to 1%, and with a pulse width variable from 0·2 μsec to 2 sec, but constant rise time. A second output is provided by a pulse which precedes that of the main output by an adjustable amount in the same range.

Cossor have a delayed pulse generator in which the delay is adjustable from 1 μsec to 11,111 μsec by decade switches. Marker 'pips' at 1, 10, 100 and 1,000 μsec intervals are provided.

In all forms of communication, the measurement of noise is of increasing importance and Wayne Kerr have produced a video noise-level meter. Basically, it comprises a video amplifier, with a bandwidth extending from 10 kc/s to 1·5, 3, 6 or 10 Mc/s, the upper limit being selected by a switch, and a mean-square detector. A calibrating oscillator is included.



E.M.I. type 1 delay line for 0 — 30 Mc/s giving 0·6 μsec delay in steps of 0·1 μsec.

The use of noise to measure intermodulation effects in multi-channel apparatus was demonstrated by the Post Office. The principle is, briefly, to apply noise equally to all channels except the one under test. The noise at the output of this channel is then a measure of the intermodulation products from the other channels.

Efforts are being made in several directions to extend the tuning range of klystrons. E.M.I. showed a reflex klystron (R 5222) using an external cavity resonator, a normal coverage being 8,400–10,100 Mc/s. By changing the resonator, however, the valve may be used from 5,000 Mc/s to 12,000 Mc/s. The power output is of the order of 60–100 mW.

Mullard have a method by which a tuning range of 500 Mc/s, with a mid-band frequency of 9,400 Mc/s, can be obtained from any low-power reflex klystron. Instead of using a mechanical plunger in a secondary cavity, this cavity is extended and provided with a ferrite ring. The effective length of the secondary cavity can be controlled by a magnetic field applied to the ferrite ring, which has the effect of altering the phase shift through the ring.

What amounts to a microwave wobbulator was demonstrated by G.E.C. It is intended for the alignment of wideband TR cells. The oscillator is mechanically tuned over a range centred on 10 cm and a linear timebase for the c.r. display is synchronized with it. The reflected wave from the test object is separated from the direct wave by a directional coupler and, after detection and amplification, it provides the Y deflection.

Phase measurements are of increasing importance in modern technique and Saunders-Roe showed an interesting phase meter covering the range of 50 c/s to 20 kc/s, with a phase range of 0–180°. The two input waves are squared in a circuit which automatically maintains a mark/space ratio of unity. The squared waves are fed to a pair of valves having a common load resistor. The resulting anode-voltage pulse depends for its width upon the overlap between the square waves and, hence, on the phase difference between the inputs. It is integrated and used to operate a voltmeter calibrated in phase angle.

Although computors are highly-specialized pieces of apparatus, in their details they function in the same way as any other piece of electronic apparatus, with one exception. They need a 'memory'. New methods of obtaining this are continually being exploited and the well-known grid of ferrite rings is one attractive way of obtaining a memory. Several examples were shown and it is essentially a low-impedance current-operated device. Plessey demonstrated an alternative which is voltage-operated. A barium titanate crystal in the form of a flat plate is used. It is coated on each side by a series of parallel conductors, the conductors on the two sides being at right angles. An elementary 'cell' is formed wherever two conductors cross. At low voltages, the response time is considerable, 1 msec or more but, at 100 V, it becomes as little as 1 μsec.

KLYSTRON CONTROL SYSTEM

By R. J. D. Reeves

(*E. K. Cole, Ltd., Malmesbury, Wilts*)

SUMMARY.—Wide-range tuning of a reflex klystron is not effected by simply manipulating one control, and therefore the embodiment of such a valve as the primary element of an automatic tuning system elevates the degree of the servo design problem. Manipulation of the controls must be constrained by consideration of oscillation strength, and this consideration constitutes a second error criterion. Accordingly, the paper introduces the concept of a 'control plane' to illustrate the klystron properties, and the automatic tuning problem is recast as one in two dimensions. One solution of this problem, to provide a.f.c. for primary radar, is described in detail.

Part 1 of the paper describes test equipment which presents the control plane on the face of a c.r. tube and maps either klystron mode areas and frequency contours or servo-trajectories on to the plane.

Part 2 describes a comprehensive a.f.c. system which takes cognizance of oscillation strength while searching for, or tracking the required frequency. A sampling technique for mode centring is introduced which affords minimum disturbance to the controls and provides a slightly better error criterion than mode peak finding. The mechanical tuning mechanism employs an automatic two-speed gear after the fashion of domestic radio receivers to provide exact and stable position setting. The separate problem of second-channel instability which is attendant upon the use of incoherent frequency discriminators has been successfully solved.

In Part 3 of the paper the complete circuit diagram is produced and the individual circuit functions described. Certain features such as the mode centre discriminator characteristic and the speed of the electronic tuning loop are the subject of analysis.

Part 1—The Cartograph

The Nature of the Klystron

THE reflex klystron is in universal use as a low-power oscillator in the centimetre wavebands, and the mechanism of its behaviour has been exhaustively discussed elsewhere[1].

To summarize, oscillation is sustained by mutual interchange of energy between a resonant cavity and a high-velocity electron beam passing through it. A negative electrode in the path of the beam reflects it through the cavity a second time, after which the electrons are collected. The electric field in the cavity modulates the velocity of the electrons as they pass through and results in electron bunching, these bunches causing further excitation of the cavity on the return journey. By suitable choice of reflector potential the beam can be made to present a negative impedance to the cavity and in this way sustain oscillation and permit a certain amount of r.f. power to be abstracted for use.

While the mechanism of it may be of considerable interest to an engineer committed to the use of such a valve, what directly concerns him is its quantitative behaviour as a circuit element; what are the controls and how are the frequency and level of oscillation related to them? Now the frequency is primarily determined by the dimensions of the resonant cavity, and some adjustment of these dimensions is assumed to be available as a tuning control. It is only necessary to pre-suppose that the condition for oscillation obtains, to overlook the whole problem of klystron control as a trivial one. But the governing condition is that the reflector shall be at a suitable potential, and the complication arises in that the optimum potential is not independent of the resonator frequency. Consequently, wide-range tuning requires joint manipulation of two controls and is not an elementary operation.

In addition to this the reflector itself constitutes a restricted tuning control, for although it is provided primarily as a means of presenting a negative conductance to the cavity, the beam admittance will, in general, be complex, and the natural frequency of the cavity can be pulled, to an extent depending on its Q. This important property raises the status of the reflector to that of the primary control in many applications, for it has just the attributes beloved by the circuit designer, namely direct control of frequency by a high-impedance electrode. A.F.C. systems employing klystrons often accept the restrictions of tuning range and general tolerances in order to maintain the somewhat artificial simplicity of using only this control.

The Control Plane

With a view to describing a more sophisticated control system, it is convenient to geometrize the problem by studying the klystron behaviour in terms of a vector space associated with the controls. As the cavity dimensions and reflector potential are considered to be the only controls, the aggregate of all possible settings is identified with a plane referred to as the control plane.

Fig. 1 is a map of the regions of oscillation in this plane, sketched from a 723/AB klystron by observing the current developed by a crystal detector. It is seen that permissible reflector

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voltages to maintain oscillation form a set of disconnected intervals. These are the characteristic modes of the klystron, the name referring to the distinct states of the electron beam which can result in oscillation. Within these intervals the power output varies continuously, as shown by the section through the control plane drawn in Fig. 2.

Examination of these sketches confirms that use of either control alone will only permit a restricted tuning range before oscillation ceases, and over this range the power output will not be constant.

Mode Mapping

To examine the modes in a quantitative manner, a simple instrument has been devised to present Fig. 1 on the face of a cathode-ray tube. The mechanical and electronic controls are associated with the X and Y directions respectively, and are scanned at different rates so as to produce a raster on the tube face which represents the control plane. The size of the raster is fixed but the amplitude of the control scanning is variable and modifies the area represented on the tube. Also the average value of the control settings can be altered to shift the origin of the display.

To delineate the mode regions the Z co-ordinate (beam modulation) is associated with oscillation strength, as indicated by detector current. That is to say, whenever the joint control setting results in oscillation a brightening waveform is applied to the tube, causing the modes to appear as bright areas in the control plane.

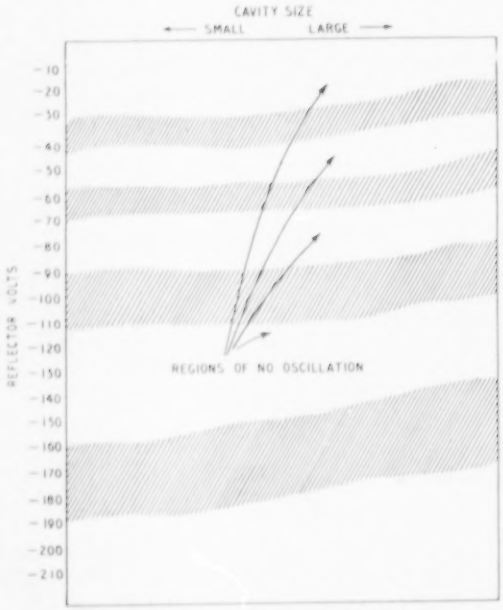

Fig. 1. *Oscillation regions in the control plane.*

In practice, there is hardly any need to have other than a fixed sweep on the cavity control, but it is very useful to be able to magnify and

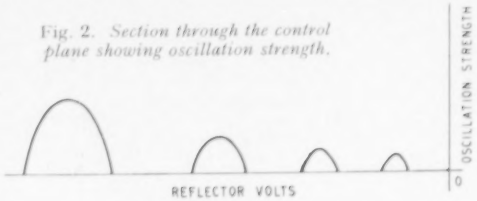

Fig. 2. *Section through the control plane showing oscillation strength.*

shift the picture in the vertical direction so that each mode can be studied individually or, alternatively, the entire set viewed together. For this reason the average (d.c.) value of reflector voltage is monitored on a meter, and vertical shift and expansion controls provided. The speed restrictions associated with mechanical controls make it necessary to explore in the X direction rather slowly, so high-speed scanning is confined to the reflector and a long afterglow screen is employed.

This map-painting instrument presents the topology of the abstract control plane in a manner reminiscent of a p.p.i. display of real space, and the name 'Cartograph' has been given to it.

Frequency Mapping and Trajectory Plotting

The mapping of crystal current only is of limited application but two interesting extensions of this basic process make the Cartograph a worth-while development tool whenever klystron behaviour has to be repeatedly examined.

First, it is a very simple matter to impress frequency contours on to the map, showing what joint control settings are necessary to generate a particular frequency. This is done by interposing an absorption cavity wavemeter between the klystron and its monitoring crystal. As a result there is a sharp notch in the transmission spectrum of the waveguide, so that the brightening waveform is interrupted whenever the klystron frequency coincides with that of the wavemeter, and a dark line is seen across the mode. The direction of these frequency contours is indicated in the sketch of Fig. 3, which reveals the restricted tuning range of the reflector control.

The second point is that the control trajectories of any klystron servo system may be plotted in the same plane.

The joint control setting corresponds to a point in the control plane and the locus of this point, as the servo performs some task, is of interest.

It is particularly useful to examine the locus in relation to the mode topology and the frequency grid, and it is practicable to do this by preceding the control plot by a few cycles of mode map painting. Then the plot of the servo trajectory can be completed before the mode background has faded from the long persistence screen, and the relation between them observed.

Cartograph Circuit

Fig. 4 is a schematic of the instrument complete with klystron and tuning mechanism, although in practice these may be integral with the servo system under observation and not attached to the Cartograph itself.

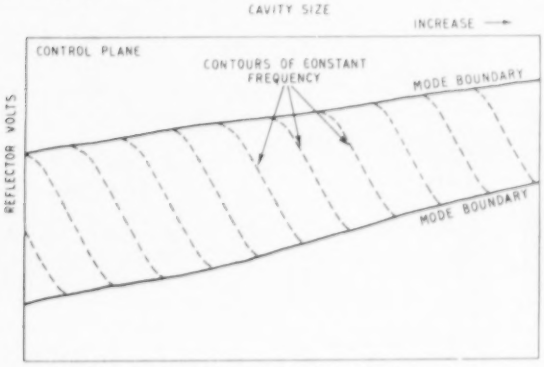

Fig. 3. *The frequency contours in the mode.*

For the map-painting process, the switches are in the positions shown and the motor scans the cavity continuously and deflects the spot in the X direction. The mean reflector potential is defined by the Y origin control and is monitored by a voltmeter, this reading corresponding to the horizontal centre line of the display. Superimposed on this potential is a sine-wave scan at a frequency of about 1 kc/s, which is also used to deflect the tube in the Y direction. The proportion of the a.c. scan that is actually fed to the reflector can be varied by the Y expansion control. The crystal-current variations resulting from scanning the reflector through several modes of oscillation are passed through a video amplifier and applied to the tube as a brightening waveform.

For the trajectory plotting feature, the relay S/3 reverses all the switches and the klystron controls are taken to the external control system. X deflection on the tube can still be performed by the cam monitor potentiometer but no corresponding facilities exist for transferring the d.c. reflector potential to the Y plates of the tube. This difficulty can be circumvented quite conveniently by leaving the same a.c. scan on the tube but now observing the instants of voltage coincidence between the scan waveform and the external reflector potential and marking them by brightening the trace. This guarantees that on the face of the tube the co-ordinates of the

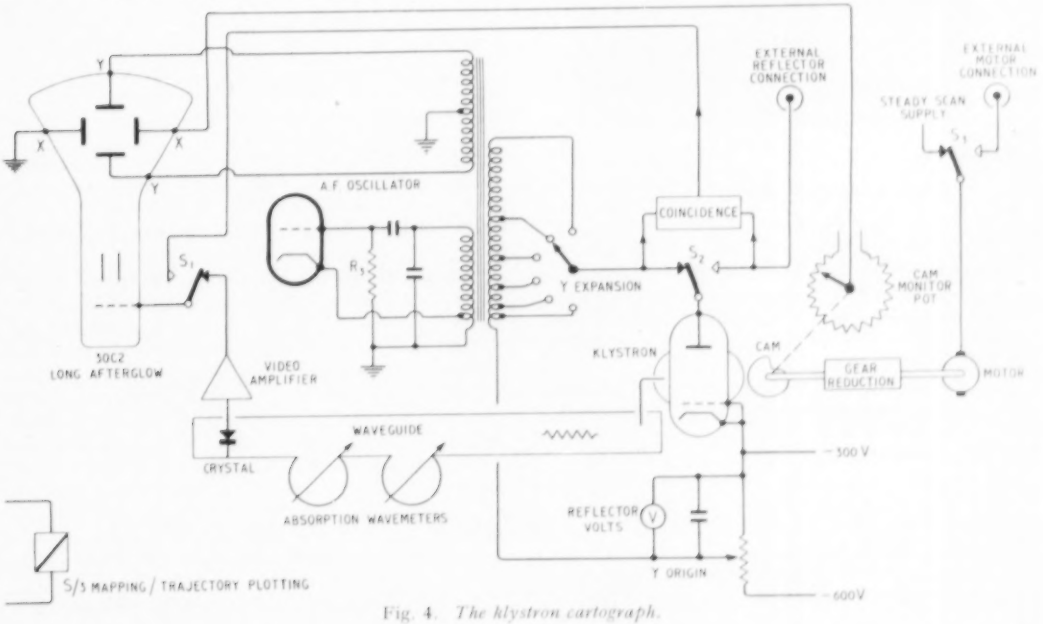

Fig. 4. *The klystron cartograph.*

trajectory are precisely the same as for the crystal current map, and the superposition of the two pictures on the long persistence screen will be a faithful presentation of the modes and servo trajectories in the ground field of control settings.

The design of the voltage-coincidence circuit presents no difficulties if it can be assumed that the scan waveform always rapidly traverses the observed reflector potential; i.e., coincidences are brief. Then a trigger pulse can be developed to fire a blocking oscillator producing a brightening pulse.

Examples of the two distinct displays produced by the Cartograph are shown in Figs. 5 and 6. The first of these maps shows one mode with a frequency contour impressed, and the second a typical control trajectory in that mode. The long-persistence screen enables these two displays to be superposed.

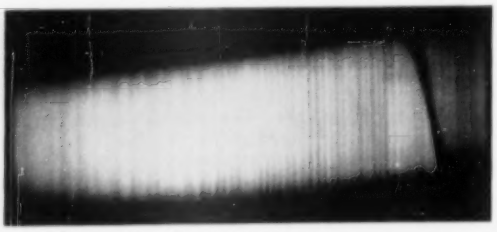

Fig. 5. *Mode map.*

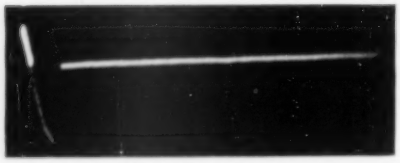

Fig. 6. *Control trajectory.*

Part 2—Comprehensive A.F.C.

The A.F.C. Problem

Where the centimetric wavebands are used for, say, radar transmission, it will be appreciated that the width of the channel spectrum is a very small percentage of the actual frequency radiated and the requirements for local-oscillator stability in the receiver are severe. However, if the degree of mistuning in the i.f. amplifier can be measured and fed-back degeneratively to the local oscillator, then the stability requirements are transferred to the error-measuring circuit at the i.f. and, at this frequency, they can easily be met. But the usual discriminators such as the Travis, Foster Seeley, Weiss, only provide a narrow region of frequency discrimination, outside of which little or no tuning error is indicated and until the i.f. beat note drifts into this region the a.f.c. loop will not be able to bring the local oscillator under control. In other words the presence of a frequency control loop alone is no guarantee of correct tuning unless means are provided to 'capture' the local oscillator at whatever frequency it initially sets itself. To make the oscillator intrinsically stable enough to stay within the capture region is virtually as stringent a requirement as keeping within tolerable tuning error without a.f.c. Inevitably the concept of a 'searching' local oscillator presents itself, whereby the oscillator is driven or cyclically drifts through a certain frequency spectrum until capture is effected. This complication, coupled with the complexity of correctly tuning a reflex klystron oscillator (as discussed in Part 1), tends to make the a.f.c. system occupy a disproportionate amount of both the radar unit and the attention of the designer.

It is a point to be decided in the initial stages of the development whether to make the a.f.c. as elementary as possible and in deference to this make the local oscillator (and transmitter) as stable as possible, or to go for completely comprehensive a.f.c. and relax the restrictions elsewhere. It is the author's contention that the comprehensive system to be described here is sufficiently compact to warrant adoption in many types of radar.

The Control System

Fig. 7 is a block schematic of the a.f.c. system referred to, which provides wide-range search and frequency control, and at the same time maintains the klystron controls at mode centre. The system is capable of locking to either of the two local oscillator channels which generate an i.f. beat, and it permits stabilization of klystron supplies to be dispensed with.

The equipment was designed for a primary radar in which a sample of the radiated signal is directly available for the frequency discriminator and there is therefore no noise problem, but this is not an essential restriction.

The output of the frequency discriminator and of the time discriminator are the source of error signals in the servo system and the Miller integrator and motor-control valves are the control points. In the block diagram the lock-on relay is shown closed and the time discriminator output (mode-centring error) is fed to the motor control, while the frequency error is fed to the reflector. Although the error signals are not coupled at all inside the control system, each error is a function of both controls, so the two loops are not independent and the system is properly described as a two-dimensional servo, which can be conveniently discussed in terms of the control plane.

In the absence of an i.f. beat the lock-on relay

drops out and diverts mode-centring error signals to the reflector, while a residual search bias keeps the motor scanning the cavity. In this condition there is no output from the frequency discriminator. The search trajectory in the control plane is therefore the line of the mode centre, whereas when the system locks on, the controls are constrained to be on one of the two frequency contours which generate an i.f. beat. The intersections of these contours with the search trajectory represent points at which both error signals are nil, and the control system can come to rest on either of them. It is assumed to be of no consequence whether the local oscillator adopts a frequency above or below that of the transmitter.

Mode Centring by Time Discrimination

The block schematic Fig. 7 is actually a plan of a working demonstration model. The waveform insets represent c.r.t. displays associated with the mode-centring process, and the relation between them is shown in Fig. 8. A single cycle of a sine wave (loosely referred to as a sinusoid) is superimposed on the reflector potential and the crystal-current variations observed; the waveforms (2) or (3) are obtained depending on whether reflector voltage or time is used as a display base. The object of doing this is to project the mode shape (2) on to a time scale where mode-centre displacement can be more easily determined. The mode mapping (3) is actually a double representation since the sinusoid traverses the mode twice, and the centre half of the sinusoid would suffice for this process but no advantage would accrue, and the waveform (1) is more easily generated. Now the sinusoid is skew symmetric about its centre point t_0, and the product of (1) and (3) will also be so if the mode map is symmetric about t_0. The average value of this product will then be zero but, if the mean reflector potential is not at mode centre, a greater part of the mode map will appear on one side of t_0 on the time scale and then the product waveform will have a d.c. content with magnitude and sign related to those of the reflector displacement. This constitutes the error signal. Fig. 8 is drawn with a small misalignment of the reflector and it is seen that the majority of the mode is projected into the second half of the mapping period. The corresponding c.r.t. displays are shown in Figs. 9 and 10.

The sinusoid actually performs a gating function in the time discriminators and not a genuine multiplication, so the skew waveform is virtually a truncated sinusoid in the demodulation process, but this is equally suitable for the purpose.

Features of the Mode Projection Process

If the sinusoid did not terminate after one cycle but persisted continuously, and if it were of very

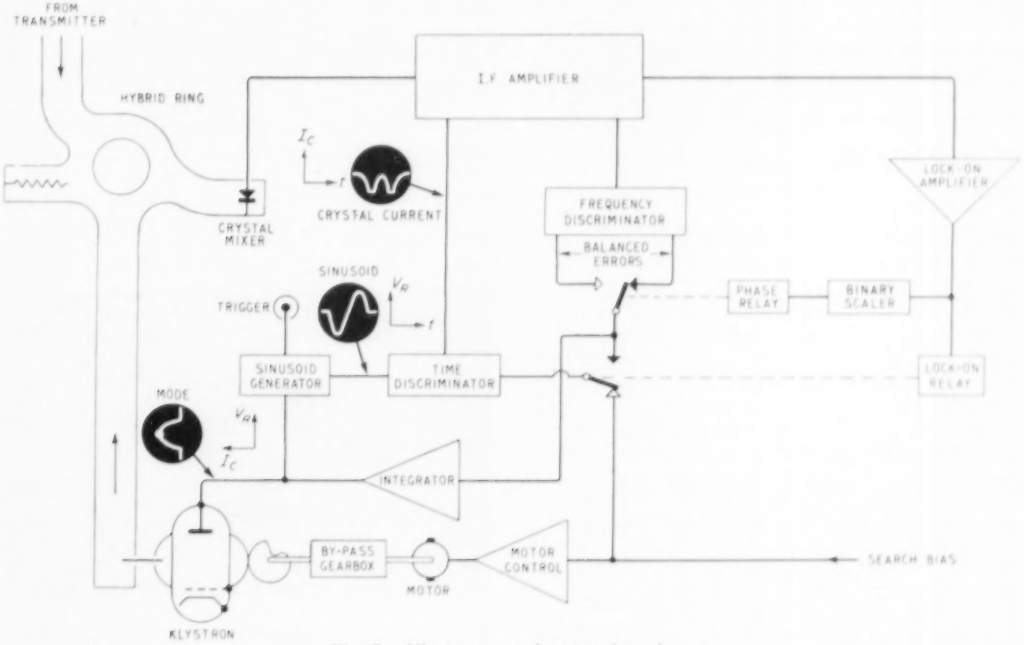

Fig. 7. *Klystron control system for a.f.c.*

small amplitude so that it explored only the immediate neighbourhood of the existing reflector setting, then the output of the time discriminator (now more correctly described as a phase-conscious rectifier) would be an indication of the magnitude and sign of the slope of the mode at that point, and the servo could be made to home on to a point of zero slope. This is the process employed in some form by all previous klystron control systems in which mode centring is attempted. In this system, however, the sinusoid amplitude is about two mode widths

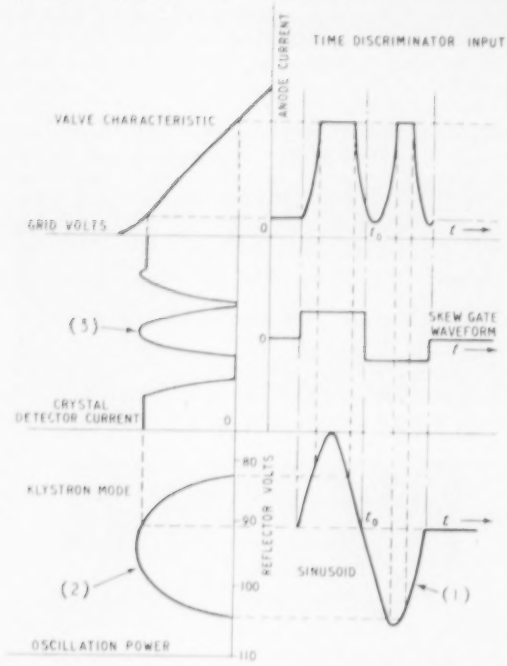

Fig. 8. *Sinusoidal projection of mode on to time scale.*

peak-to-peak and between mapping times there are no perturbations on the reflector from this source. The mapping times can be very short and can be distributed arbitrarily so that, although reception has to be abandoned completely at these moments, this restriction is more apparent than real. The mode-centring loop becomes in fact a sampling servo and the necessary rate of sampling depends on the highest error frequency required to be handled, which is likely to be very low indeed. The sampling rate could therefore be correspondingly low but, in practice, there is usually a dead time in the radar at the end of every repetition period and it is convenient to initiate the sinusoid excursion there. Alternatively, if the duty cycle of the sinusoid is low,

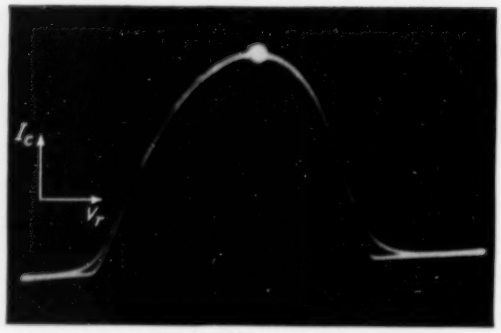

Fig. 9. *Klystron mode.*

say, less than 0·01, then it may be run asynchronously with the transmitter and that proportion of received signals lost.

The use of a large-amplitude sinusoid confers two advantages which characterize this control system, the most significant one being the improved mode pull-in performance. Since error signals are developed even if just the peak of the sinusoid is encountering the mode, the reflector will be pulled in to the mode centre. Using a sinusoid which is two mode widths peak-to-peak, the pull-in range may be three times greater than a system which requires the mean reflector potential to be actually on the mode before centring occurs. The capture range, which is so important at the time of switching on, has therefore been usefully extended.

The other feature is that peak power is no longer the servo target but rather the discriminator measures displacement from the centroid of (weighted) mode area. If the mode is symmetrical no difference in rest position results but, if the mode is appreciably asymmetrical, the reflector will adopt a position from which it is better able to cope with transient disturbances in either direction. This would seem to be a more useful criterion than that of peak power.

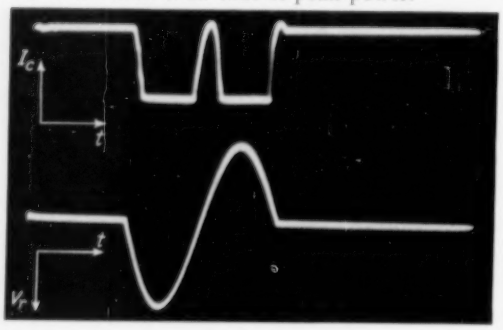

Fig. 10. *Sinusoid deviation and consequent crystal current variations.*

Frequency Discriminator

In principle, the frequency discriminator is quite conventional but the design is governed by the fact that the i.f. is not continuously available but is sampled when the transmitter fires. The result is that either a clamp or an integrator is required to store the output during the space period, and also the signal incoming to the discriminator is not a spectrum line but a frequency distribution of sin x/x form. Therefore, the overall discrimination characteristic is not the frequency characteristic of the discriminator itself but the product of the discriminator characteristic and the incoming pulse spectrum as a function of the misalignment of the two. In this situation it does not follow that a small peak separation in the discriminator corresponds to high error sensitivity for, if the pulse energy is distributed over a wide band, the discriminator characteristic tends to be swamped. Moreover, unless the peak separation is comparable to the width of the principal part of the pulse spectrum it is quite possible for the a.f.c. to lock to a minor peak.

In this particular design a half-microsecond pulse has to be catered for, and a peak separation of 5 Mc/s is used in the discriminator.

A buffer amplifier is interposed between the discriminator and the pulse-stretching circuits in order that the design of the discriminator is not complicated by excessive loading effects. The error output is available in either polarity according to the state of a phase relay, so that ambiguity of sign arising from ignorance of the phase of the i.f. can be resolved. A separate d.c. output is developed whenever a signal appears in the discriminator, and this constitutes a lock-on signal.

Lock-on Procedure

The presence of the lock-on signal indicates that the search has brought the local oscillator close enough to the required frequency to be put under the control of the frequency discriminator and the change-over from searching to frequency tracking is effected by a lock-on relay. In this design, the chief function of this relay is to divert the mode-centring errors from the reflector to the motor circuit and swamp the search bias with mode-error signals. The reflector is left to handle the frequency-error signals which appear at the same time as the lock-on voltage.

The search–follow transition is therefore well defined, and this fact enables a simple relay scaling circuit to be employed to combat a type of instability which threatens all automatic systems in which both first and second channel frequencies fall within the swept spectrum. In a conventional (incoherent) frequency discriminator, increments of local-oscillator frequency are preserved in magnitude but are ambiguous in sign and, consequently, a control loop which is designed to stabilize at the i.f. on one channel will evade it on the other. Such evasion throws the pulse spectrum to the edge of the i.f. channel, at which point the lock-on signal disappears and the reflector control recentres itself in the mode in accordance with the normal search condition. But this, of course, is the same control setting as that which originally brought about the lock-on, so the unstable cycle repeats indefinitely. However, by relating a loop-phase reversing relay to the lock-on relay by a scale of two, successive lock-on operations find alternate loop phases existing and, if the first lock-on is not stable, the second one will be. In this way the control system will settle the klystron on the frequency above or below that of the magnetron indiscriminately, depending which is first encountered in the search cycle. This procedure not only circumvents the difficulty of ignoring the unstable channel during the search but is likely to shorten the searching time.

Discussion of the homing trajectory in the control plane is deferred until the gearbox features have been described.

Mechanical Features

The choice of tuning motor and the design of its associated mechanism are of central importance for this unit may well govern the bulk, complexity and reliability of the whole equipment. Although there is no appreciable load to consider, precise position control is required and this is one reason why speed control of the motor would be desirable. But a mechanical system which incorporates local tachometer feedback is likely to be bulky and demand a substantial power output from the control circuit. Simplicity is best maintained by departing from linear methods and using a constant-speed motor with on-off control by a polarized relay. Unfortunately, a gear reduction that results in acceptable search is not compatible with that required for fine tuning, and it is necessary to correlate a gear change with the lock-on relay so that both fast search and a stable control loop may be realized. The way that this is achieved here is a development of a device featured for many years in domestic radio sets, and depends on having a unidirectional search (with a quick return or flashback). The full gear reduction for low-speed tuning is available only for a limited interval of movement, beyond which a section of the reduction is by-passed and coarse tuning provided. But the low gear interval travels with the by-pass drive and can always be entered by reversing the motor. The effect is that the motor approaches

its final rest position at high speed and overshoots. The error signals reverse the motor which then enters the low-gear interval and returns to the rest position at greatly reduced speed. This interval has to be at least as wide as the initial overshoot but, apart from that, the extent of it is not important. Twice the overshoot is a reasonable size to adopt. Continuous tracking in one direction would, of course, cause a momentary change into high gear periodically, but this is tolerable and in any case is not anticipated as typical.

The two tuning speeds are chosen quite independently of each other according to the search speed requirements on the one hand and the maximum desirable loop gain on the other, and the transition from high speed to low is performed at the moment of reversal when the mechanism is stationary, so the homing approach is essentially smooth. Changing from low speed to high normally occurs shortly after unlocking

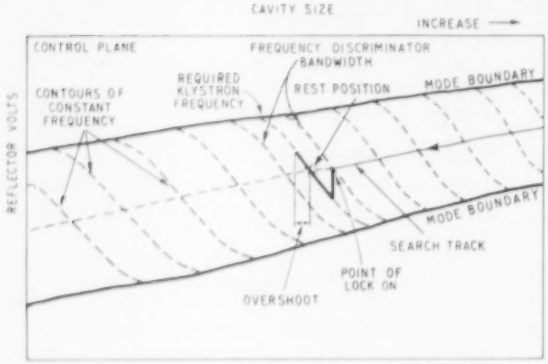

Fig. 11. *The homing trajectory.*

when the servo has started searching and the load then suffers a velocity step. In fly-power position where this impact is not severe, the by-pass type of gear box has many applications because of its simplicity both in construction and in functioning without any auxiliary information being required. It would, for instance, be equally applicable in a search–follow system in which the lock-on transition is not well defined by a relay operation.

Unless the search were unidirectional the mechanism would have to traverse the low-gear interval at each end of the scan and this would be a time-consuming and pointless process. The difficulty is avoided by having a flash-back mechanism such that the tuning piston is returned to the start from the end of every search scan. This, in fact, is the simplest search procedure, for the mechanism must not be reversed by local switching because of the loop phase ambiguity which then arises when the system locks on. A two-way search requires limit switches which reverse the search bias in the control unit, with the attendant hazard in the

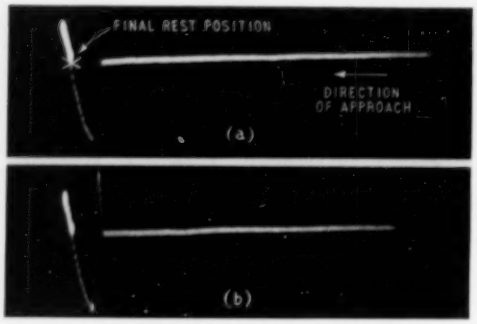

Fig. 12. *Approach with loop phase; (a) correct, (b) incorrect.*

assumption that the control system is functioning as soon as the motor is switched on.

The present quick-return mechanism is an entirely self-contained unit which can be run independently of the control system for the purpose of testing, and of course in conjunction with many different types of control system.

The Homing Trajectory

The behaviour of the complete system is best described with reference to the homing trajectory, shown in Fig. 11. When the equipment is first switched on, searching starts as soon as the control system warms up, and mode pull-in occurs as soon as the klystron warms up and crystal current is available. Thereafter, the search trajectory is the locus of mode centre, and centring is maintained even during the mechanical flash back. This condition persists for a few minutes before the transmitter comes on, after which it is possible to develop an i.f. beat when the klystron is brought to an appropriate frequency. The two channels are relatively close so, if the klystron is ascending the frequency spectrum, it is more likely to encounter the low channel first and lock on to that. A lock-on voltage is developed as soon as the search trajectory encounters the frequency line corresponding to the edge of the frequency-discriminator band, and immediately the reflector is put under the control of the discriminator. The subsequent motion depends upon whether the frequency-control loop phase is correct or not; if it is incorrect, the reflector will drive away from the required frequency contour (upwards in Fig. 11)

until the lock-on signal disappears, and then re-set itself on the search trajectory. Immediate relocking occurs with the phase now corrected. The speed of the reflector control is fast compared with that of the piston so that these motions appear as a vertical movement in the sketch.

With the loop phase correct the reflector jumps to the required frequency contour and from this moment the i.f. is held. There is at this point, however, a mode-centring error, so the motor (supplied with this error signal) continues to turn in the same direction to annul the error. The reflector loop constrains the joint control setting to be always on the frequency line, which is therefore the new trajectory of the motion. The intersection of this with the search trajectory is the point where both error signals are zero and therefore the final rest position, but on the first approach the motor will overshoot because the mechanical by-pass is still in operation. The first reversal, however, engages the final gear reduction and the controls move back to the rest position with adequate loop damping. Fig. 12 (a) is a photograph of the homing trajectory when the loop phase is correct for the channel that is first encountered, whereas Fig. 12 (b) shows the adapting transient undergone when the phase is unsuitable on the approach.

If the transmitter signal disappears, the servo will unlock and start to travel along the search trajectory from its rest position, but motion will be slow at first until the low gear interval has been traversed, and this is advantageous in that, if the interruption is only momentary, the system will relock and return instead of escaping into an unnecessary search cycle.

(To be continued)

LINE TRANSMISSION CIRCUITS

Effects of Longitudinal Voltage and Current

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(Engineering Division, B.B.C.)

SUMMARY.—This article describes in some detail the means by which longitudinal currents and voltages are introduced into transmission circuits and associated apparatus. The effects of these currents and voltages, and methods by which these effects may be eliminated or reduced, are discussed.

IT is necessary to start by defining four terms which will be used frequently:—

1. A 'Longitudinal Current' flowing in each wire of a pair of wires may be defined as half the algebraic sum of the instantaneous currents flowing in the wires at a given point.
2. A 'Circulating Current' flowing in each wire of a pair of wires may be defined as half the algebraic difference of the instantaneous currents flowing in the wires at a given point.
3. A 'Longitudinal Voltage' is defined as half the algebraic sum of the instantaneous voltages between each wire of a pair of wires and a third conductor at the given point. (The third conductor is usually earth.)
4. A 'Circulating Voltage' between the two wires of a pair may be defined as the algebraic difference between the voltages between each wire and a third conductor (usually earth) or alternatively as the product of the current flowing through, and the magnitude of the impedance between the pair at that point.

If, in order to determine the magnitude of the circulating voltage at a given point, an impedance is connected across the pair, then the magnitude of the 'open-circuit' voltage may be determined by Thévenin's theorem.

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Consider the general case of current flow in the light of (1) and (2) above. Any current flowing may be represented as (a) a circulating current, I_c, flowing in opposite directions in the two wires, and (b) a longitudinal current, I_0 flowing in the same direction in the two wires. The currents I_c and I_0 may consist of the instantaneous sum of several waveforms and, further, each may consist of a wanted and an unwanted component.

The general case of voltage, as in (3) and (4) above may be represented by an equipotential voltage, V_0, between each wire and the third conductor and a voltage, V_a, between one wire and the third conductor equal in magnitude but opposite in sign to a voltage V_b between the other wire and the third conductor. The longitudinal voltage V_0 and the circulating voltage $(V_a—V_b)$ may each consist of the instantaneous sum of several waveforms.

In general, therefore, a pair of wires may have a wanted signal voltage between them and, in addition, an unwanted signal voltage between them, also a wanted and an unwanted longitudinal voltage may be present between each wire and a third conductor.

It will be seen that the concepts of longitudinal current and voltage are closely linked. It is

convenient, however, to have established them both, as some problems involving their consideration may be solved more easily by one concept than the other.

It will now be shown that if circulating current is present with longitudinal current then, in general, a longitudinal voltage exists.

Fig. 1 represents a small length ΔL of a pair of wires (A and B) distance L from a generator connected to A and B, the primary constants of the pair being uniform over the length ΔL.

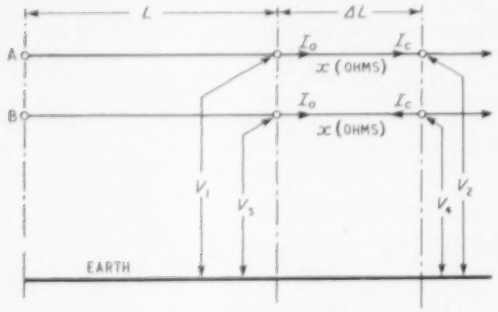

Fig. 1. *Pair of wires A and B near earth.*

Let I_c be the circulating current in the section, the resistance of the section of wire be x ohms, (neglecting inductance, capacitance and leakance) and let I_0 be the longitudinal current present. Further, let voltage V_1 to earth and V_3 to earth (at the entrance to ΔL.) be identical but opposite in sign; i.e., by definition, at this point no longitudinal voltage exists. Consider now the voltages V_2 and V_4 to earth (at the end of ΔL.)

$$V_2 = V_1 - (I_c + I_0)\,x \quad \ldots \quad \ldots \quad (1)$$

$$V_4 = V_3 - (I_0 - I_c)\,x \quad \ldots \quad \ldots \quad (2)$$

Now $V_3 = -V_1$

Hence $V_4 = -V_1 - (I_0 - I_c)\,x$

$$= -V_1 + (I_c - I_0)\,x \quad \ldots \quad (3)$$

From (3) $$-V_4 = V_1 - (y_c - I_0)\,x \quad (4)$$

Comparing (1) and (4) it will be seen that $-V_4 \neq V_2$ and hence at $L + \Delta L$ a longitudinal voltage exists.

Let us consider how longitudinals may be introduced into a pair of wires. Fig. 2 shows how this occurs by electromagnetic coupling either with the pair itself or with a circuit element connected to the pair.

Referring to Fig. 2(a) G is a conductor adjacent to the pair AA', BB' and is carrying a current I_g in the direction shown. The flux lines f_1, f_2, etc. due to I_g cutting AA' and BB' produce currents I_{0a} and I_{0b} of magnitude increasing with I_g and decreasing with d_1 and d_2 the distance of AA' and BB' respectively from G. It will be seen, therefore, that if the distance d between AA' and BB' is small with respect to d_1, then $d_2 \rightarrow d_1$ and therefore $I_{0a} \rightarrow I_{0b}$. Further, if AA' and BB' are transposed along their length, this also tends to make $I_{0a} = I_{0b}$.

Clearly, for current to flow along A'A and B'B there must be a return path A to A' and B to B'. In general this is via the admittance to earth from A and A', similarly from B and B'.

Referring to Fig. 2(b) assuming that T_1 is a perfect transformer balanced to earth, and that Y_a and Y_b are the admittances to earth at A and B of the pair AA', BB' and that adjacent to T_1 is a source of electromagnetic flux S. The flux linkages f_1, f_2, etc. from S to the secondary of T_1 are such that the induced voltage per turn progressively decreases along the length of the secondary winding. Consider C, the centre point of the winding having admittance Y_c to earth. The induced voltages from C are V_1 to A and V_2 to B; and clearly $V_1 \neq V_2$.

For a normal pair of wires in good condition, Y_a is approximately equal to Y_b and, therefore, the voltages at A and B to earth are approximately equal to V_1 and V_2 and as $V_1 \neq V_2$, by definition longitudinal voltages exist at A B due to induction from S. It will be seen that, in the cases where electromagnetic introduction of logitudinals occur, the source impedance is low compared with the path followed by the induced currents via earth admittances. Hence, it is convenient to consider such induction as emanating from a constant-voltage source.

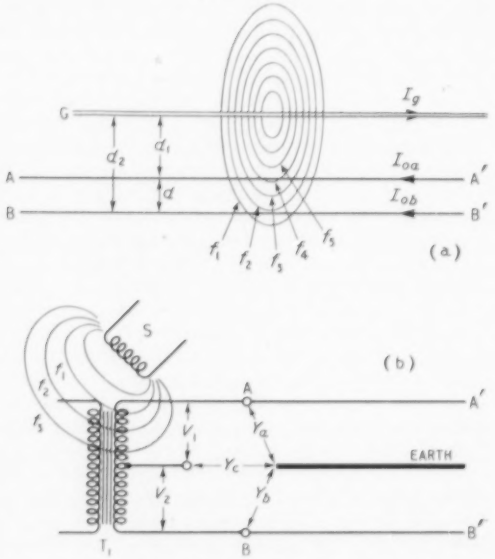

Fig. 2. *Pair of wires near a conductor G carrying current (a) and wires fed from a balanced transformer (b).*

Let us now consider the introduction of longitudinals by electrostatic coupling, either with the pair itself or with a circuit element connected to the pair.

Fig. 3 (a) shows a conductor G adjacent to pair A B and having a potential V with respect to earth and negligible current flow along its

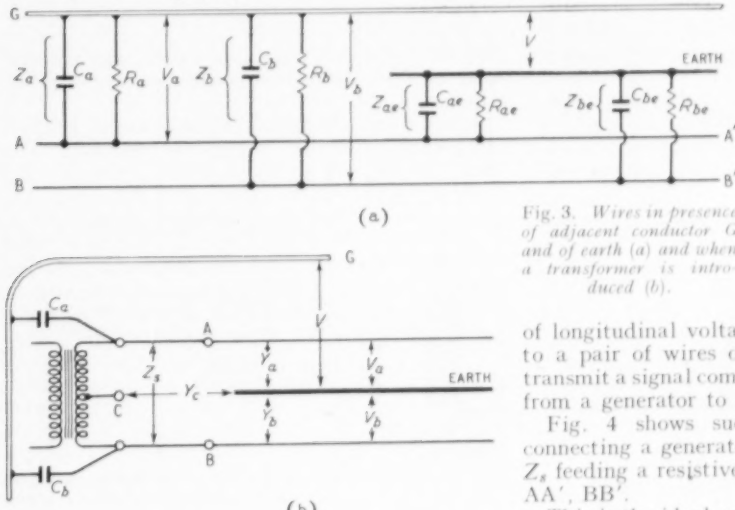

Fig. 3. *Wires in presence of adjacent conductor G and of earth (a) and when a transformer is introduced (b).*

length. Consider the section of A B chosen to have impedance Z_a between A and G, Z_b between B and G, Z_{ae} and Z_{be} to earth respectively.

Consider the voltages V_a and V_b from G to A and B

$$V_a = V\frac{Z_{ae}}{Z_a + Z_{ae}}$$

$$V_b = V\frac{Z_{be}}{Z_b + Z_{be}}$$

It will be seen that V_a and V_b, though possibly of equal magnitude, will be of the same sign and, therefore, by definition a longitudinal voltage exists at this point. Further if $Z_a \neq Z_b$ and/or $Z_{ae} \neq Z_{be}$ then $V_a \neq V_b$, the conditions of an induced unwanted signal voltage.

Fig. 3 (b) shows the same transformer T_1 as in Fig. 2 (b) connected to pair A B. In this case, however, a conductor G is adjacent to T_1 having voltage V with respect to earth and capacitance C_a and C_b to the ends of the secondary winding as shown.

Assume again admittance Y_a, Y_b and Y_c and further, that the secondary-winding impedance Z_s is equally divided at C.

Due to V a voltage drop V_a will exist across Y_a which is now part of the potential divider $\frac{1}{j\omega C_a} + \left\{\frac{1}{Y_a} \text{ in parallel with } \frac{1}{Y_c} + \frac{Z_s}{2}\right\}$ and, similarly, a voltage drop V_b will exist across Y_b and will be of the same sign as V_a with respect to earth. Therefore by definition a longitudinal voltage exists.

The electrostatic introduction of longitudinals is, in the main, by the capacitance (leakance normally being negligible) between the pair of wires (or apparatus connected to it) and the generating source.

In general, this capacitive susceptance will be very much smaller than the earth admittances. It is convenient, therefore, to consider such inductions as emanating from a constant current source.

We will now consider the practical implication of longitudinal voltages and current in relation to a pair of wires over which it is desired to transmit a signal comprising a complex waveform from a generator to a distant load.

Fig. 4 shows such a transmission circuit connecting a generator with internal impedance Z_s feeding a resistive load R via a pair of wires AA', BB'.

This is the ideal case and the rest of this paper shows how far practical circuits fall short of this ideal and some methods that may be adopted to overcome errors introduced by longitudinals.

Fig. 5 shows the circuit of a pair AA'—BB' connecting a generator G to a load L. (G incorporates an output transformer T_1 with

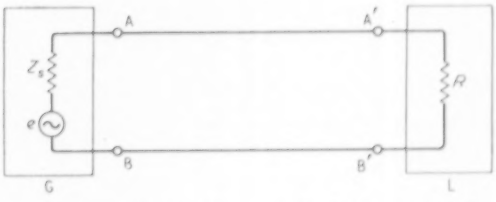

Fig. 4. *General form of load L connected to a generator G.*

generator E_a in the primary and L is the first stage of a triode amplifier with an input transformer T_2.) The pair of wires has imposed upon it, due to an external coupling, a longitudinal current I_0, an unwanted circulating current I_u, and due to the e.m.f. of generator G (produced by I_p in the primary T_1), the wanted signal current I_a.

The current I_u flowing through T_2 will produce an unwanted signal at the grid of the valve. I_0 may be represented to have been produced by a

generator E_0 of e.m.f. e connected to earth via an impedance Z_0 to the centre point of the secondary winding of T_1. We may conveniently represent the inter-winding capacitance of T_2 as lumped in C_1 and C_2 at either end of the windings.

The return path to earth at A′ B′ is via the impedance Z_a and Z_b to earth of the input terminals of L (other paths to earth, of course, exist along the pair). Consider impedance Z_b; this is equal to $1/j\omega C_2$ (the parallel leakance being neglected).

The impedance Z_a (neglecting minor quantities) is made up of a parallel path $(j\omega L_p + 1/j\omega C_2)$ and $(j\omega L_s + 1/j\omega C_1)$ where L_p and L_s are the primary and secondary shunt inductance.

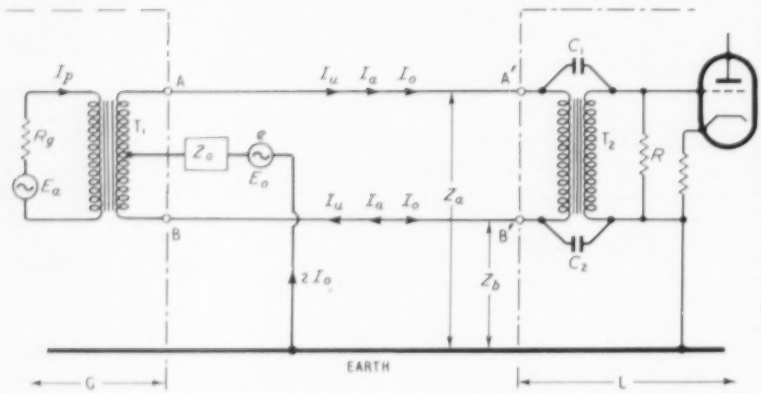

Fig. 5. *Line connected to generator and valve amplifier through transformers.*

Clearly, $Z_a \neq Z_b$ and therefore the voltages across them due to I_0 flowing through them are not equal. Hence, due to I_0, a further potential difference is established across the input terminals of the load which will cause a further unwanted voltage at the grid of the valve.

It will be seen, therefore, that even if I_u is not present, I_0 will cause an unwanted voltage in the load.

A similar analysis will reveal that an equivalent state of affairs exists at the generator end.

Hence, it may be said that a longitudinal current (or voltage, by a similar analysis) induced in a transmission pair will cause an unwanted signal to appear at the grid of a valve amplifier connected to that pair. The effect would be even more serious if no input transformer were present where in general $Z_b = 0$ and Z_a is of the order of one megohm.

Obviously the magnitude of the unwanted signal will depend upon the value of I_0 and the difference between Z_a and Z_b. (Similar considerations apply when determining the unwanted signal added to the generator output.)

It is impossible completely to isolate a pair of wires and the apparatus connected to it from other current or voltage-carrying conductors. Therefore, to keep unwanted signals to a minimum—apart from screening and twisting the pair to reduce the magnitude of induced longitudinals—it is necessary to make the apparatus terminal-to-earth impedances balanced to the best possible limits.

This can be achieved to very good limits by the use of the 'balanced line transformer' or 'repeating coil'.

The essential features of these are an earthed electrostatic screen interposed between the primary and secondary windings; and the construction of the primary windings in two halves. These are made in such a manner that if the windings are connected series-aiding (for circulating current) the potential gradient with respect to the junction, in each winding, is virtually the same when a longitudinal voltage with respect to earth is applied across their outer terminals.

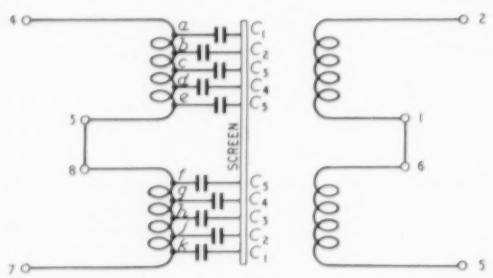

Fig. 6. *Arrangement of windings and capacitance distribution in transformer.*

Fig. 6 shows the capacitance distribution in such a transformer.

The 'line windings' 4-3, 8-7 have equal capacitances C_1, C_2, C_3, etc. as shown. Now between points 'a' and 'b' the resistance, inductance and capacitance are identical with those between points 'j' and 'k', et seq, until points 'e' and 'f'; let us consider the passage of longitudinal current through the line winding. As shown before, the path of such a current entering 4 and 7 is through the earth admittances at 4 and 7.

Due to the equal distribution of primary constants in windings 4-3, 8-7, it will be seen that points 'e' and 'f' will be equipotential with respect to earth and the current flow through each related turn will be identical in magnitude but opposite in direction, therefore, the resultant magnetic flux will be zero; i.e., no voltage will be induced in the secondary winding due to the longitudinal current flowing in the primary. Capacitance will of course exist between the screen and windings 2-1, 6-5, but the balance of these windings is not, in general, critical, since in most cases any unbalances here will be swamped by the unbalances of following apparatus. On certain occasions, however, it may be desirable to have these windings balanced, and consideration is given in the design of such a transformer to achieve this as far as possible.

Careful manufacturing methods can produce a very good approximation to balance and such transformers enable quite large longitudinal currents to be harmlessly by-passed from the signal path.

Reverting to Fig. 5, consider the effect of introducing a 'repeating coil' between A'B' and the load L. I_0 will now cause no signal at the grid of the valve but I_u, the unwanted signal current, will be unaffected and will still develop a voltage at its grid. If we suppose that I_u and I_0 are induced in the generator G, then the insertion of a repeating coil between it and AB will remove I_0 leaving I_u. Therefore, although the line transformer can be extremely effective in removing the deleterious effect of longitudinals, an unwanted circulating current will always cause an unwanted voltage at the load.

Let us now consider the passage of a longitudinal current along a pair of wires having relative variation in some primary constants. Fig. 7 shows a section PP'—NN' of a pair AA'—BB' with longitudinal current I_0 entering the section (and at PN the longitudinal voltage $= V_0$). The section having resistance $R + x$ from P — P' and R from N—N' and leakance to earth from P' $= 1/R$ and from N' $= 1/(R' + x')$. Suppose $x = 0$, then voltage P' to earth and N' to earth $= V_0 - I_0R$.

Current to earth from P'

$$= \frac{V_0 - I_0R}{R}$$

and from N'

$$= \frac{V_0 - I_0R}{R' + x'}$$

Let current leaving section at P'

$$= I_p = I_0 - \frac{V_0 - I_0R}{R'}$$

Let current leaving section at N'

$$= I_m = I_0 - \frac{V_0 - I_0R}{R' + x'}$$

Therefore $I_m > I_p$

Hence the longitudinal current at this point

$$= \frac{I_m + I_p}{2}$$

There also exists due to I_0 a circulating current

$$I_c = \frac{I_m - I_p}{2}$$

Therefore due to unequal leakances the longitudinal current has produced a circulating current and, by a similar analysis, if x is finite a similar result ensues. Hence, it will be seen that any irregularity of primary constants in a transmission pair will cause unwanted signal currents from any induced longitudinal. Thus, the capacitance and leakage to earth, also the wire resistance and inductance, ought to be identical for each wire along the length of the pair.

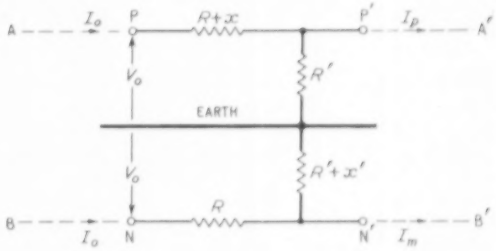

Fig. 7. *Conditions in a line in which the primary constants vary.*

Take the case where the transmission pair is one of many such made up as a cable. It was shown earlier that to reduce to a minimum the value of circulating currents magnetically induced, the wire-to-wire spacing should be small and the pair continuously transposed along its length. The second precaution is easy to realize, but the first, although desirable from the consideration of space, must not be carried so far as to make the wire-to-wire capacitance of the pair too great (about 0·07 μF/mile is about the maximum permissible for cable transmission of audio frequencies).

By care in manufacture all these requirements can be closely met in each section produced. The jointing of sections to produce one long cable must be done so as not to produce resistance unbalance in the pair at the joint, and it may be necessary to transpose the pairs of the section to maintain the best capacitive balance of a number of sections. It will also be seen that if the insulation of a cable breaks down, the wire-to-

earth leakance of a pair may be seriously unbalanced. These unbalance conditions also have to be avoided at any other connection points in the overall circuit such as the 'jumpering' in a distribution frame, fuses, jacks, etc.

The remaining pairs of wire in a cable must be considered as potential inducers of unwanted signals in the pair under consideration. Fig. 8 shows a pair AB with an adjacent pair CD carrying a signal current I_d and having capacitances C_a, C_b, as illustrated.

Let us assume I_d to be small in magnitude (which will generally be the case), then due to this current in C and D being in opposite directions and the distance between wires C and D being small the magnetic field will be negligible and it will induce no appreciable current in A and B.

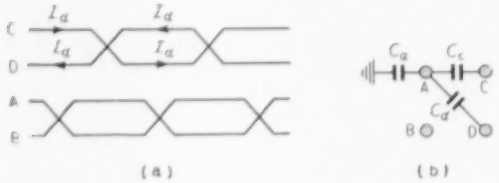

Fig. 8. *Two pairs of adjacent transposed wires (a) and the relevant capacitances (b).*

In Fig. 8 (b) is shown a section of AB and CD, let V be voltage of C to earth and $-V$ be the voltage of D to earth due to a signal voltage of $2V$ between C and D and let A have a resultant voltage V_a to earth.

$$V_a = V\frac{C_c}{C_a + C_c} - V\frac{C_d}{C_a + C_d}$$

Now if $C_c \neq C_d$, $V_a \neq 0$. Similarly, a voltage V_b from B to C and D are not equal.

Hence it is essential in such a cable to ensure the best wire-to-wire capacitance balance for each pair between the pairs, and between each pair and the cable sheath (earth), as mentioned above. It will be obvious that one way of reducing the effects of unbalance capacitance coupling is to enclose each pair of wires in an electrostatic screen. This of course increases the bulk of the cable and also its cost. It further means that the pairs cannot be quadded. Screening, however, plays a valuable part in reducing induction in individual pairs of wires not part of a metallic sheathed cable—whose very sheath provides overall screening from outside sources of interference.

Let us consider the virtues of applying an earth to the individual screens in an individually-screened pair cable. The screens can only conveniently be earthed at the ends of the cable. It must be obvious that if there exists in a given pair a longitudinal current (with respect to earth) then, if there is a low resistance path to earth surrounding the pair, the current flowing through the capacitances between the screen and the pair will all return via the screen to the distant earth.

Such a current flowing in a conductor adjacent to other screens and pairs will magnetically induce unwanted currents in them. Therefore, in general, it is most inadvisable to make any earth connection to the screens of individual pairs.

In order to provide an output which is balanced with respect to earth, and isolated from it, it is usual to feed the output from a valve to its load via a transformer. Owing to practical design limitations, however, direct capacitance coupling between the primary and secondary windings of such a transformer cannot be completely eliminated. The result of this remanent coupling is the generation of longitudinal voltage in the output circuit.

In Fig. 9 let the direct capacitive coupling be represented as a lumped capacitance C between the top and bottom ends of the windings as shown. (For simplicity they are shown as having equal values, if they were not it will be seen that then the unbalance will be accentuated.) Also, let Z_1 and Z_2 be the impedances to earth from 1 and 2.

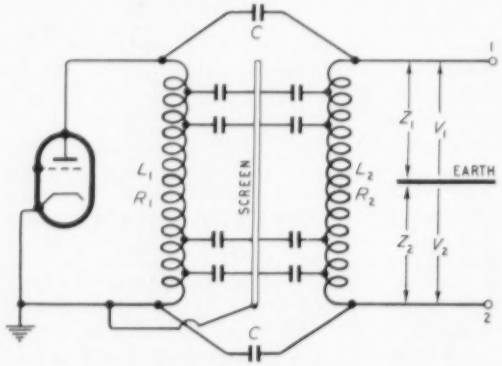

Fig. 9. *Illustrating direct capacitive coupling through a transformer.*

Z_1 = The parallel combination of $R_2 + j\omega L_2 + 1/j\omega C$ and $R_1 + j\omega L_1 + 1/j\omega C$

$Z_2 = 2/j\omega C$ (neglecting the high impedance path in parallel) see below.

(Where R_1 and L_1 are the resistance and shunt inductance of the primary, and R_2 and L_2 those of the secondary.)

In practice (for a 600-Ω output impedance)

such a transformer will have L_1 and R_1 much greater than L_2 and R_2 thus

$$Z_1 = R_2 + j\omega L_2 + 1/j\omega C$$

and therefore $Z_1 > Z_2$. If the generated voltage appearing across 1 and 2 is V, the voltage between earth and 1 is V_1, and the voltage between earth and 2 is V_2

$$\text{Then } V_1 = \frac{-VZ_1}{Z_1 + Z_2}$$

$$\text{and } V_2 = \frac{VZ_2}{Z_1 + Z_2}$$

and as $Z_1 \neq Z_2$, $V_1 \neq -V_2$.

Hence, a longitudinal voltage exists at 1 and 2 with respect to earth.

The source of these voltages may be represented by a generator having an impedance $1/j\omega C$ (the real part of the impedance may be neglected) in series with a point on the transformer secondary and earth, and having a frequency equal to and a waveform related to that of the source of circulating current. The actual point on the winding is where the impedances to earth are both equal and are $(Z_1 + Z_2)/2$.

If $Z_1 = Z_2$ then there will of course be no longitudinal voltage induced due to unbalanced earth admittances. Longitudinal current may, however, be present due to another type of coupling; e.g., external. In these cases the disturbing generator may be represented as acting at the centre point of the winding.

In parenthesis, it will be seen that these longitudinal voltages exist despite the presence of the earthed electrostatic screen and, therefore, it might appear likely that similar conditions would apply to the repeating coil. This is in fact so, but to a much smaller degree because the direct winding to winding capacitance of a repeating coil is only of the order of 6% of that for an average input or output transformer (typically 15 pF instead of 250 pF.) This is achieved by having a transformer of approximately unity turns ratio, requiring an inductance that need only be large compared with the impedance of the apparatus connected to it—normally of the order of 600 Ω and therefore the limited number of turns required enables the most efficient use to be made of the available winding space to result in minimum inter-winding capacitance.

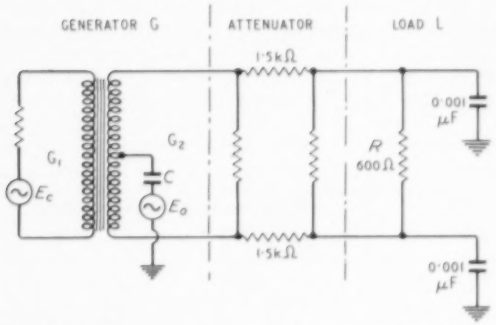

Fig. 10. *Behaviour of attenuator in presence of longitudinal current.*

Consider the apparatus illustrated in Fig. 10, where an attenuator having 20 dB loss is inserted between the generator G having a source of circulating e.m.f. E_c which produces circulating voltage V_c across the load resistor R, and a source of longitudinal e.m.f. E_0 connected through an impedance $1/j\omega C$. If the circulating source impedance is 600 Ω and equal to R and the attenuator give 20 dB loss under these conditions, the total value of series resistance in the pad will be approximately 3,000 Ω.

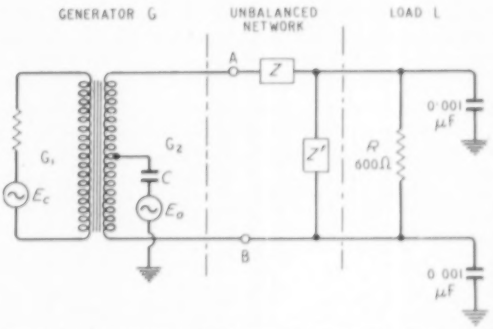

Fig. 11. *Behaviour of unbalanced network in presence of longitudinal current.*

Let us further suppose that the earth admittances at the load terminals are mainly capacitive (a typical value is of the order of 0·001 μF and the value of C in the generator may be taken as 250 pF). Let us consider the behaviour of the attenuator in respect of the two types of current flowing through it. If the frequency of G_1 is 10 kc/s, then the frequency G_2 is also 10 kc/s. The impedance represented by $1/j\omega C$ is approximately 64,000 Ω, while the load to earth impedances are approximately 16,000 Ω. The attenuation of flow of the longitudinal current by introducing 1,500 Ω in an effective total path of 144,000 Ω is therefore negligible (less than 0·2 dB). The shunt arms of the attenuator, of course, introduce no loss at all to this current. It is obvious, therefore, that the 20-dB attenuator has increased the proportion of longitudinal to circulating voltage at the load terminals by nearly 20 dB or 10 times. At 1 kc/s the generator G_2 impedance and earth admittances are correspondingly increased and the effect more marked,

in that the longitudinal loss is decreased; but, of course, the amount of this current will be considerably less.

Let us now consider Fig. 11 which shows an unbalance introduced in series with the generator G and load L of the previous example. Let the voltage developed across the load due to G_1 be V. As has been shown earlier, the presence of the unbalance will cause an unwanted circulating voltage V_c at the load terminals due to any longitudinal voltage applied at points A and B. As the generator G_2 impedance is capacitive the amount of longitudinal current flowing through the unbalance will tend to increase with frequency and hence the unwanted voltage developed across the load will also tend to increase with frequency. As the generator G_2 will have a frequency identical to that of G_1 and a waveform related to that produced by G_1, it will be seen that V_c may add or subtract from the wanted circulation voltage V at the load terminals and, as voltage V_c will vary with frequency, the effective insertion loss of the unbalance element introduced will not be that calculated by consideration of its loss to the circulating current, due to G_1 and this deviation is frequency dependent.

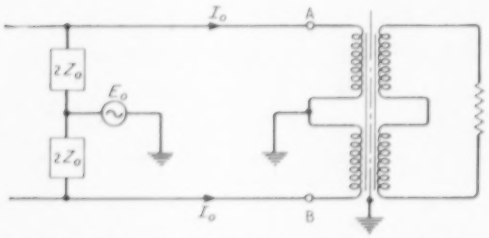

Fig. 12. *Effect of earthing centre tap of transformer at the receiving end.*

An example of such a connection is an amplifier feeding a transmission line followed by an equalizer terminated by an amplifier. There will be, in general, a longitudinal current due to imperfection in the sending-amplifier output transformer. This current will cause an unbalanced equalizer to have other than its designed result at the input to the receiver amplifier. Secondly, if in the transmission line a longitudinal current is induced from an external source, the equalizer unbalance will cause a 'noise' (or unwanted signal) voltage at the receiver amplifier output. It becomes necessary, therefore, to eliminate both these efforts by inserting a repeating coil immediately before the unbalanced equalizer.

The effects described when considering Fig. 10 show that the introduction of transmission loss following an output transformer considerably worsens its effective longitudinal output compared with the circulating voltage it is designed to deliver at the attenuated output. This use of a variable output attenuator is frequently resorted to in test oscillators and the above effect must be taken into account when using such apparatus.

Lastly, let us consider (Fig. 12) the receiving end of a transmission line terminated with a repeating coil, and the effect on longitudinal current due to applying an earth to the centre tap of the line winding on the transformer. Let I_0 be the longitudinal current flowing due to generator E_0. We have seen earlier that depending on whether I_0 is caused by electromagnetic or electrostatic induction the longitudinal generator approximates respectively a constant-voltage or constant-current generator.

In the first instance the connection of the centre tap will considerably lower the earth impedance from A and B, hence I_0 will increase and the greater part of it will flow through the windings.

If a constant-current generator is assumed, connection of the centre-tap earth will not increase I_0, but will increase that portion of it flowing through the windings. As this concept postulates Z_0 infinitely great with respect to all the other impedances in the return path this seldom occurs in practice, but it is useful to adhere to this concept when considering the earth-connected centre point.

If the current flowing through the line windings increases, then any resistive or inductance unbalance they may have will cause a larger induced voltage in the secondary windings.

In the case of electromagnetically-induced longitudinals the increase in I_0 due to the earth connection will cause a corresponding increase of unwanted circulating voltage at the load, due to any unbalances before it. The earth connection in this case will therefore considerably worsen the interference due to unbalance.

This fact provides a simple test for a practical circuit to determine how unwanted noise voltages are being induced. If the induction is electrostatic, the earth connection will not normally much increase the unwanted signal.

In general, therefore, it will be seen that it is not desirable to apply an earth connection to the centre tap of a repeating coil, used to terminate a transmission line.

For some other types of transformer which have fairly small but quite unbalanced winding to earth capacitances, the earth connection may improve the signal-to-noise ratio in the presence of longitudinals, but this necessitates reasonable inductive and resistive balance in the 'line winding' of such a transformer.

AMPLITUDE LIMITATION IN *LC*-OSCILLATORS

Using Two Biased Diodes

By Ziya Akçasu

(Assistant Professor at the Technical University of Istanbul, Turkey)

SUMMARY.—An *LC*-oscillator which includes two biased diodes in the grid circuit to limit oscillation amplitude is analysed. Amplitude of oscillation and harmonic distortion are calculated and represented by curves. From these curves it is observed that oscillations can be modulated linearly; no frequency modulation occurs during amplitude modulation and harmonic distortion is very low. Also the circuit is found to be adequate for high frequency stability. The distinctive feature of the circuit is shown to be that the static conditions of the valve, such as bias voltage and direct anode current, are independent of the oscillation amplitude. A practical circuit is described.

1. Introduction

In general, an *LC*-oscillator is composed of an oscillation circuit which determines oscillation frequency and of a negative resistance which supplies energy to maintain oscillations. An oscillation circuit can be reduced to a parallel combination of an ideal inductance, a capacitance and a positive resistance which represents the total losses in the real circuit. Maintained oscillations occur when the energy supplied by the negative resistance becomes equal to that absorbed in the positive one. Oscillation amplitude is limited, therefore, by the non-linearity of either of these resistances[2]. Thermistors[5] (negative temperature-coefficient resistors) and Varistors[8] (voltage-dependent resistors) are the well-known non-linear positive resistances. They can be connected in parallel with the oscillation circuit to limit oscillation amplitude[4].

A negative resistance is usually obtained by means of a valve operated as a positive feedback amplifier. Hence, the non-linearity of a negative resistance may be due either to the non-linearity of the valve characteristic or to the non-linearity of the feedback circuit. An example of the former case is the conventional *LC*-oscillator, where the oscillation amplitude is limited by the non-linearity of the valve characteristic; the valve is operated in these oscillators as a class C power amplifier[9]. The latter case, where the non-linearity is due to the feedback circuit, is the subject of this paper and will be dealt with in detail.

2. Description of the Circuit

Fig. 1 shows the block diagram of the oscillator circuit; it consists of a tuned amplifier and a positive feedback circuit. The amplifier comprises a valve which is operated in class A. The valve characteristics are assumed to be linear. This property of the circuit is one of its distinctive features.

Feedback is obtained through an attenuator which provides the coupling and the proper phase shift. The attenuator is followed by two biased diodes*. When the output voltage of the attenuator is below the bias voltage E, the diodes are not conducting and hence the attenuator is terminated by the non-conducting resistances of the diodes; the attenuation factor of the attenuator is then denoted by β. If, on the other hand, the bias voltage is exceeded the diodes become conducting and, since their conducting resistance is assumed to be negligible, the output voltage remains equal to the bias voltage E. Hence a clipped waveform is obtained. The ratio of the fundamental component of this wave to that of the input voltage is defined as the attenuation factor. It will be shown below that the attenuation factor in this new situation depends upon the magnitude of input voltage. Thus, one obtains a non-linear feedback. It is important to note that the diodes are not operated as rectifier elements producing a negative bias to control oscillation

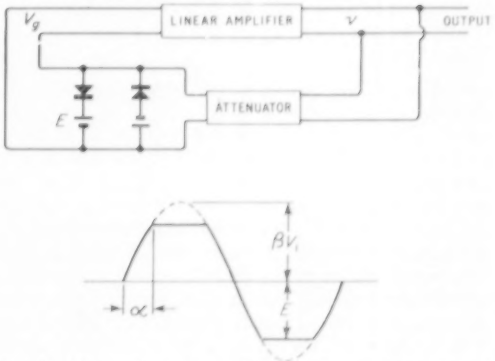

Fig. 1. *Basic form of oscillator and waveform clipped at the level E by the diodes.*

MS accepted by the Editor, June 1955

* See Appendix.

amplitude. This point is also one of its distinctive features[7]†.

It is now assumed that the input impedance of the attenuator is so much greater than the output impedance of the amplifier that the former can be neglected since they are in parallel. This assumption is made in order to prevent the non-linear feedback circuit impairing the linearity of the amplifier.

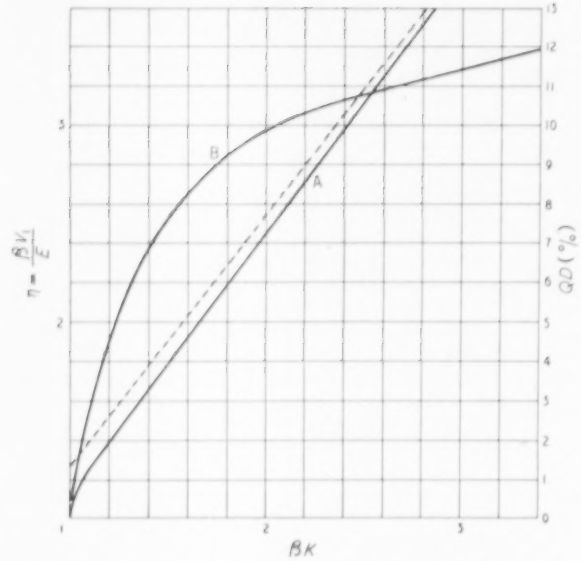

Fig. 2. *Relation between loop gain βK and clipping level (curve A) and product of distortion and Q (curve B).*

3. Amplitude of Oscillation

It will be shown below that the output voltage v of the oscillator contains only odd harmonics and its Fourier expansion is of the form:

$$v = V_1 \sin \omega t + V_3 \sin 3\omega t + \ldots \quad \ldots \quad (1)$$

As a first approximation, the effect of harmonics on the amplitude will be neglected. This is equivalent to assuming that the voltage applied to the diodes is a pure sine wave and is equal to $\beta V_1 \sin \omega t$ (Fig. 1). This assumption makes it possible to express the clipped sine wave at the grid as a simple Fourier series. When the voltage applied to the diodes is an arbitrary periodic function of time, the output voltage will be again composed of only odd sine harmonics; but the expressions giving the amplitudes of harmonics will now involve Bessel functions[3]. Hence the calculations are very long even if only the third harmonic is taken into account.

† See Ref. 9, p. 205.

The Fourier expansion of a clipped sine wave, as shown in Fig. 1, is known to be

$$v_g = B_1 \sin \omega t + B_3 \sin 3\omega t + \ldots \quad \ldots \quad (2)$$

where

$$B_1 = \frac{2}{\pi} \beta V_1 (\alpha + \frac{1}{2} \sin 2\alpha) \quad \ldots \quad \ldots \quad (3)$$

$$B_3 = \frac{1}{3\pi} \beta V_1 (\sin 2\alpha + \frac{1}{2} \sin 4\alpha) \ldots \quad \ldots \quad (4)$$

In these expressions α is the angle shown in Fig. 1 and defined as

$$\sin \alpha = \frac{E}{\beta V_1} \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

This expansion holds as long as $\alpha < \pi/2$; i.e., $\beta V_1 > E$. When $\beta V_1 < E$, the diodes are not conducting and hence the oscillator circuit is linear. In linear circuits the condition of self-excitation is known to be

$$\beta K \geqslant 1$$

where K denotes the gain of the amplifier at the resonant frequency. If the above condition is satisfied, oscillations start spontaneously and the oscillation amplitude increases exponentially until the instantaneous value of the voltage applied to the diodes reaches the bias voltage. From now on the voltage peaks are clipped and the circuit becomes non-linear. Thus, harmonics are produced and expansion (2) holds.

When the steady state is reached the following relation holds between the fundamental components of the input and output voltages of the amplifier:

$$B_1 K = V_1 \quad \ldots \quad \ldots \quad \ldots \quad (6)$$

By substituting B_1 from Equ. (3) one obtains

$$\frac{2}{\pi} (\alpha + \frac{1}{2} \sin 2\alpha) = \frac{1}{\beta K} \quad \ldots \quad \ldots \quad (7)$$

If the attenuation β and the gain K are given, α can be obtained from this equation; oscillation amplitude V_1 is then calculated from Equ. (5). The curve A in Fig. 2 shows the relation between βK and η where

$$\eta = \frac{V_1 \beta}{E} \quad \ldots \quad \ldots \quad \ldots \quad (8)$$

The curve approaches asymptotically the dotted straight line $\eta = \frac{4}{\pi} (\beta K)$.

From the graph one can draw the following conclusions:

(a) If the attenuation β and the bias voltage E are kept constant, oscillation amplitude increases with the gain of the amplifier, being roughly proportional to it.

(b) Since for a given value of βK, η has a certain value, the amplitude of the oscillation V_1 is proportional to the bias voltage E. Therefore, the control of the bias voltage of the diodes enables the oscillation amplitude to be varied within wide limits. In an actual circuit the extent of the linear portion of the valve characteristic sets an upper limit to the oscillation amplitude.

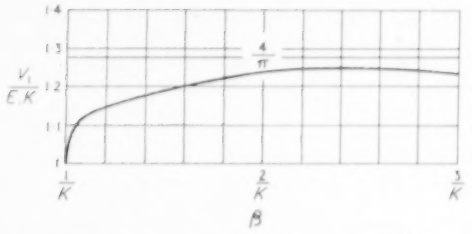

Fig. 3. *Variation of amplitude of oscillation with attenuation.*

It also follows that the oscillation can be modulated by superposing an audio-frequency voltage upon the bias voltage E. Since the amplitude is exactly proportional to the bias voltage, the modulation characteristic will be linear, provided the operation remains in the linear part of the valve characteristic. The depth of modulation can be increased up to 100%. As is shown below, no frequency drift is present during this modulation.

In class C LC-oscillators, however, the oscillation is modulated by varying the anode supply voltage but, when the anode voltage falls to very low values during modulation, the valve ceases to oscillate, which sets a limit for the depth of modulation. Besides this, the modulation characteristic is not linear in such oscillators and frequency drift occurs during modulation.

(c) When the gain K and the bias voltage E are held constant, the oscillation amplitude varies with the attenuation β as shown in Fig. 3. It is observed from the curve that for $\beta = 1/K$ (i.e., when the oscillation just starts) the oscillation amplitude is equal to EK. For higher values of β the curve first increases rapidly, then reaches a very flat maximum and falls slightly. For still higher value of β the curve approaches asymptotically to $4/\pi = 1{\cdot}27$; this part of the curve is not shown in the figure. It follows that the magnitude of feedback denoted by β has not much effect on the oscillation amplitude, but the values of β corresponding to the flat part of the curve are advisable since, on this part of the curve, any small variation of β causes a relatively smaller change in amplitude. It will be shown in what follows that the harmonic distortion increases with β, and suitable values of β are between $1{\cdot}1/K$ and $1{\cdot}3/K$.

4. Harmonic Distortion

Fourier expansion of the grid voltage given in Equ. (2) shows that the magnitude of the harmonics decreases rapidly with increasing harmonic number. Moreover, the gain of the tuned amplifier falls steeply for the higher harmonics, and only the third harmonic need be considered in the calculation of harmonic distortion.

The impedance of the tuned circuit for the third harmonic is

$$|Z_3| = \frac{3}{8}\omega L \qquad \ldots \quad \ldots \quad \ldots \quad (9)$$

where L is the inductance of the tuned circuit. If g_m denotes the mutual conductance of the valve, the gain of the amplifier for the third harmonic is

$$K_3 = \frac{3}{8}\omega L g_m \qquad \ldots \quad \ldots \quad \ldots \quad (10)$$

where the anode resistance of the valve is assumed to be much greater than the impedance of the tuned circuit. Since, for reasons which will be mentioned later, screen-grid valves are preferred for this oscillator circuit, this assumption is justified.

By remembering $K = \omega L Q g_m$ one can transform Equ. (10) into the form

$$K_3 = \frac{3}{8}\cdot\frac{K}{Q} \qquad \ldots \quad \ldots \quad \ldots \quad (11)$$

where Q is the Q-factor of the tuned circuit.

The amplitude of the third harmonic at the grid is given in Equ. (4). By multiplying this by K_3 one obtains the amplitude V_3 of third-harmonic at the anode. Dividing by V_1 gives the harmonic distortion, i.e.,

$$D = \frac{V_3}{V_1} = \frac{1}{8\pi}\cdot\frac{\beta K}{Q}\left(\sin 2\alpha + \frac{1}{2}\sin 4\alpha\right) \quad \ldots \quad (12)$$

where α should be substituted from Equ. (7). By combining Equs. (7) and (12) one can see that the product DQ depends solely upon βK. The curve B in Fig. 2 represents the relationship between DQ and βK. The curve approaches asymptotically to the value of 0·125 for the higher values of βK.

Equ. (12) and the curve show that:—

(a) Harmonic distortion is inversely proportional to the Q-factor of the tuned circuit. A high Q-factor is essential therefore for low harmonic distortion. One of the causes which reduce the effective Q-factor of the tuned circuit in an oscillator is the internal resistance of the valve. Besides the reasons which will be mentioned later, screen-grid valves are preferable to others because

they have a high anode resistance. Alternatively, current feedback can be used to increase the internal resistance, which also serves to improve the linearity of the valve characteristic.

(b) Harmonic distortion increases with the product βK and it has been shown previously that a suitable value of βK is 1·2. The harmonic distortion corresponding to this value is found from the curve to be $(4{\cdot}6/Q)\%$ which approximately gives 0·1% for a tuned circuit having $Q = 50$. This shows that the harmonic distortion can be easily decreased to very low values.

(c) The harmonic distortion is independent of the oscillation amplitude. This point will be returned to later.

5. Frequency of Oscillation

The well-known interdependence between the frequency of self-maintained oscillations and the harmonic content is expressed mathematically as follows[1,2,*]

$$\frac{\Delta\omega}{\omega_0} = -\frac{1}{2}\left[3\left(\frac{V_2}{V_1}\right)^2 + 8\left(\frac{V_3}{V_1}\right)^2 + 15\left(\frac{V_4}{V_1}\right)^2 + \ldots\right] \quad \ldots \quad (13)$$

where ω_0 is the natural frequency of the tuned circuit and where $\Delta\omega = \omega - \omega_0$. If only the third harmonic is considered, this equation takes the form

$$\frac{\Delta\omega}{\omega_0} = -4D^2 \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

This suggests that a harmonic distortion of 0·1% gives a frequency reduction of about four parts in a million.

The fact that harmonic distortion is independent of oscillation amplitude together with Equ. (14), shows that frequency reduction is also independent of oscillation amplitude. Hence, it follows that amplitude variations due to amplitude modulation do not give rise to any frequency drift. This point has been previously referred to.

In a class C oscillator, however, amplitude modulation is always accompanied by some degree of frequency modulation. This can be seen by considering the kind of amplitude limitation in these oscillators. The rectification at the grid produces a negative bias which makes the operating point more negative when oscillation amplitude increases. Thus, the character of the non-linearity of the valve and, consequently the harmonic content, change with the oscillation amplitude. This explains why frequency modulation occurs during amplitude modulation. Variations of anode voltage due to modulation also cause the operating point to shift. The above argument is also true for the circuit given in Ref. 7 where an external diode is used to produce the negative bias voltage.

It has been found that the frequency variation due to change in interelectrode capacitance with variation of operating conditions of the valve is of far more significance than effects due to the harmonic content[9]. The capacitance between any two electrodes in a valve is dependent on the space charge as well as on their geometrical disposition. Therefore it varies with the voltage applied to the electrodes and the mean anode current. Since the valve in the described oscillator is operated in class A and its characteristics are assumed to be linear, the mean anode current does not vary with the oscillation amplitude, and it follows that the capacitance introduced in the tuned circuit is independent of the oscillation amplitude; consequently, no frequency modulation is present during amplitude modulation.

This argument shows that a valve having a small anode–earth capacitance should be preferred; this is one of the reasons for using a pentode for the oscillator circuit discussed in this paper.

In pentodes, the variation of the direct current with change of anode-supply voltage is small; the same is also true for the mutual conductance of a pentode. If the valve is automatically biased by a cathode resistor, this variation is still smaller, because any variation in anode voltage is accompanied by an opposite change in the grid voltage. Hence, the variations of anode voltage have little effect on the frequency stability of the oscillator.

Variation of the screen-grid voltage is more significant than that of the anode voltage, therefore it is sufficient to stabilize only this voltage.

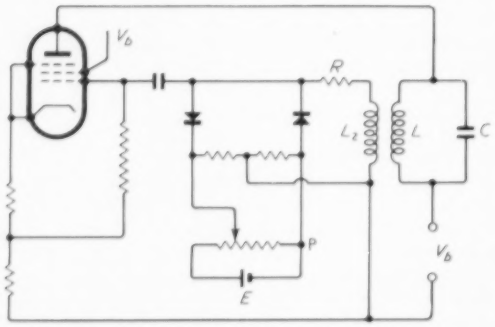

Fig. 4. *Circuit employed for experimental work.*

6. Application

The circuit used by the author for the experiments is shown in Fig. 4. Germanium crystals were used as diodes. The potential divider P enables the bias voltages on the two diodes to be

*See also Ref. 9, p. 70.

varied simultaneously. The resistor R is inserted in series with L_2 in order to prevent the diodes short-circuiting the coil L_2. This ensures that the tuned circuit is not damped by the feedback circuit. The parts played by the other elements in the circuit are self-explanatory.

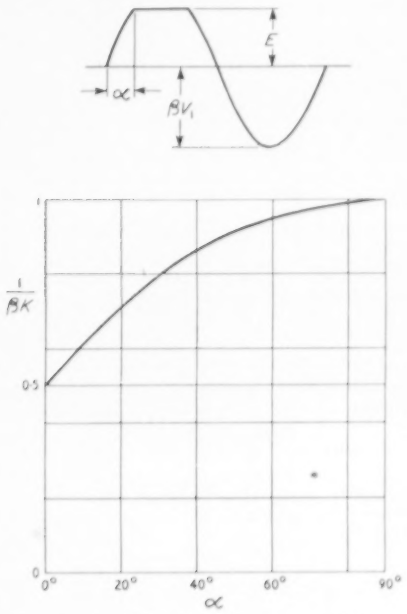

Fig. 5. *Conditions for a single diode clipper.*

7. Conclusion

The use of diodes as the non-linear element makes it possible to operate the valve in class A and also to assume its characteristics to be linear. The distinctive feature of the circuit lies in the fact that diodes are not used to produce a negative bias by rectifying the oscillations; they simply produce a clipped wave whose fundamental is not proportional to that of the voltage applied to them. Since the operating conditions of the valve do not change with the oscillation amplitude, the circuit has a high stability of frequency with a linear amplitude modulation, without any frequency drift and finally produces oscillations with a very low harmonic content.

APPENDIX

If only one diode is used at the grid, amplitude limitation is not always possible; beyond a certain value of coupling, oscillation amplitude is not under control and increases indefinitely. To show this one can follow the same procedure as that used in the case of double diodes. The clipped sine wave in this case has the form shown in Fig. 5 and its Fourier expansion is as follows:

$$v_g = B_1 \sin \omega t + B_3 \sin 3\omega t + \ldots + A_0 + A_2 \cos 2\omega t + \ldots$$

The fundamental component B_1 is

$$B_1 = \frac{1}{\pi}\left(\alpha + \frac{1}{2}\sin 2\alpha + \frac{\pi}{2}\right)V_1\beta$$

The equation for determining oscillation amplitude can be found as in the case of two diodes; viz.:

$$\frac{1}{\beta K} = \frac{1}{\pi}\left(\alpha + \frac{1}{2}\sin 2\alpha + \frac{\pi}{2}\right)$$

$$\sin \alpha = \frac{E}{V_1\beta}$$

The curve shown in Fig. 5 represents the relationship between βK and α. It is observed from this curve that the amplitude of oscillation has a finite value if βK has values satisfying

$$1 < \beta K < 2$$

As βK approaches 2, oscillation amplitude increases indefinitely. For values of βK larger than 2, oscillations are unlimited and never reach any steady state; the amplitude is not under control and is independent of the bias voltage E.

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The authors are professors at the École Nationale Supérieure, but at present the former is Director of the Laboratoire Central des Télécommunications, while the latter is head of the "Câbles Hertziens" Department of the French Thomson-Houston Co. The book has been specially written for those students who, having obtained a general knowledge of radiotelegraphy, wish to concentrate on the three branches of the subject indicated in the title. It is therefore divided into three parts, the first of which, after an introductory review of the fundamental laws of electromagnetism, has a chapter on the TE_{01} wave of the rectangular guide, followed by a chapter entitled "a general study of guided waves". The second part entitled "circuits" consists of four chapters dealing with dipoles, quadripoles, multipoles, and cavities. The third part entitled "aerials" is divided into two chapters, the first on the general properties of aerials, and the second on the principal types of aerial, with an appendix on Babinet's Principle. The references given at the end of each chapter are divided into general works and particular references, a résumé of the former being also given at the end of the book, together with an index and a very comprehensive table of contents.

Each chapter concludes with a number of exercises, each marked with a symbol to indicate whether it is regarded as easy, or of medium difficulty, or difficult. The preface concludes with a tribute to Maxwell "from whose celebrated equations written in 1867 all this work flows". As the authors say in the preface, the subjects dealt with in the book are no longer of interest only to specialists in radar and telecommunications, since the same techniques have penetrated into the field of nuclear energy and the particle accelerator and into other fields. The book will undoubtedly prove valuable to those with the necessary fundamental knowledge who wish to master the intricacies of waveguides and the various elements of short-wave circuits and aerials.

The names of English and German books in the bibliography have not been carefully checked. We have noted Microwave Cricuits, Principales of Microcrowave Circuits, and Ausgewhälte Fragen, but these are minor details compared with the general excellence of the book.

G.W.O.H.

Mathematische Methoden in der Hochfrequenztechnik

By Klaus Pöschl. Pp. 331 + viii with 165 illustrations. Springer-Verlag, Reichpietschufer 20, Berlin, W.35. Price Dm. 36.

Dr. Pöschl is in the development department of the Siemens and Halske valve works in Munich, and this book is the outcome of the desire to have at hand in a single volume the mathematical material which is essential to the solution of the theoretical problems of high-frequency research and development. The book is intended both for the practising engineer or physicist and for the student, and can be regarded either as a textbook or as a book of reference.

The author regards it as made up of two parts: the first ten chapters develop the mathematics and the last five deal with its application to field problems, cavity resonators, waveguides, aerials, magnetrons, waves in space-charges, etc. The early chapters deal with scalar and vector fields, determinants and matrices, complex quantities and circle diagrams, and analytical multivalued functions, conformal representation, Fourier series and integrals, Laplace transformation, statistics, linear differential equations of the second order, cylindrical and spherical functions, methods of approximate solution using step by step methods. This selection of a few headings will give an idea of the wide ground covered.

The material has been very carefully prepared and arranged: a slip giving eleven errata is inserted in the book, but unfortunately one of the corrections is in need of further correction. In addition to the table of contents, there are extensive lists of references both to books and to papers on the subject, which, it should be emphasized, is not high-frequency engineering but the application of mathematics in the problems that arise.

G.W.O.H.

CORRECTION

We have been requested to point out that in the Rola Celestion advertisement (April issue) the size of the pole piece was incorrectly given as 1/5th". This should, of course, have read 1·5".

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(Communication from the National Physical Laboratory)

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1	+0·4	N.M.
2	+0·4	0
3	+0·4	−3
4	+0·4	+1
5	+0·4	−3
6	+0·3	+2
7	+0·4	−2
8	+0·4	−3
9	+0·4	−3
10	+0·3	−3
11	+0·4	−3
12	+0·4	−3
13	+0·4	−2
14	N.M.	−3
15	N.M.	+2
16	+0.4	0
17	+0·5	0
18	+0·5	N.M.
19	+0·4	0
20	+0·4	+2
21	+0·4	0
22	+0·4	−2
23	+0·5	−2
24	+0·5	−2
25	+0·5	−2
26	+0·5	0
27	+0·6	−1
28	N.M.	−1
29	N.M.	0
30	+0·6	+4

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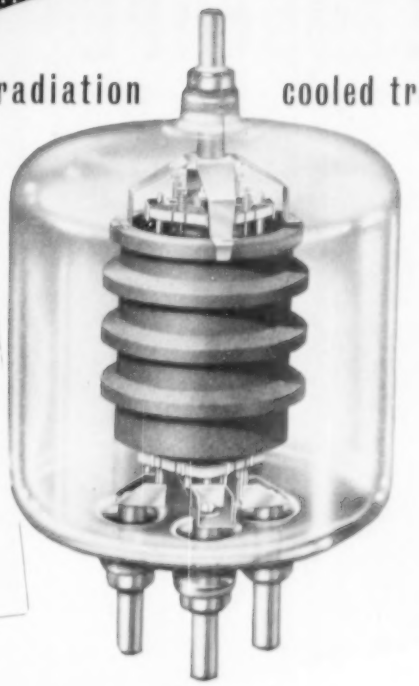

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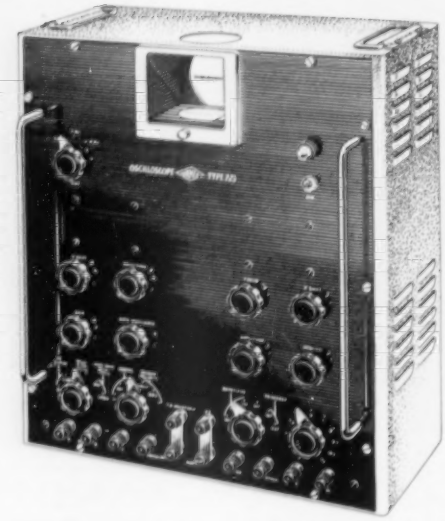

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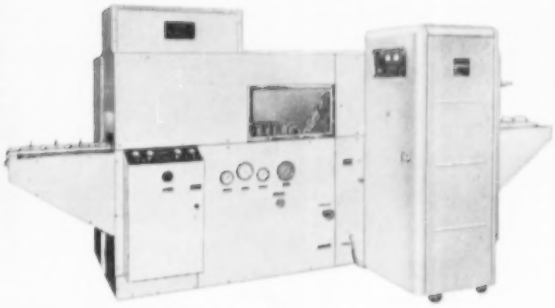

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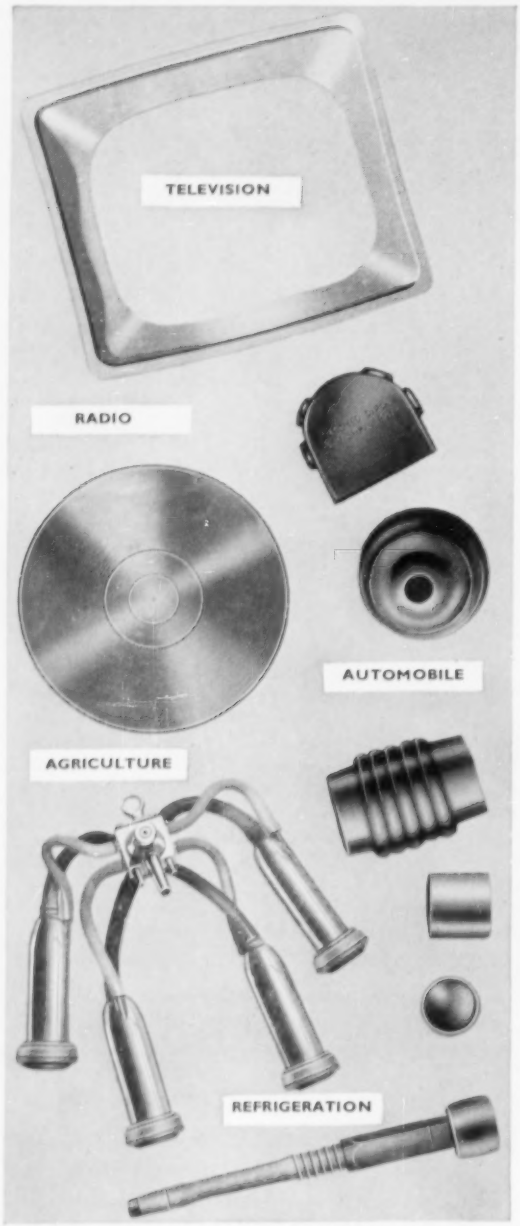

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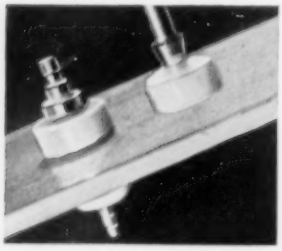

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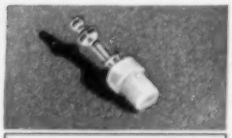

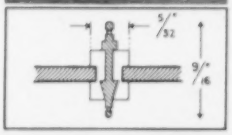

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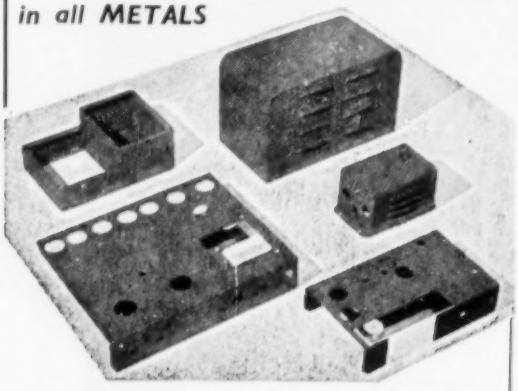

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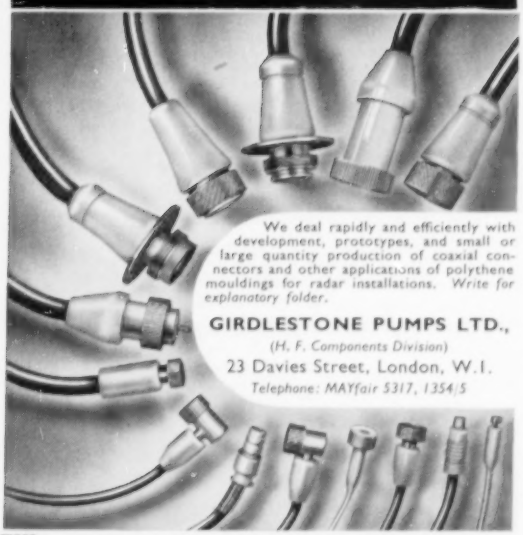

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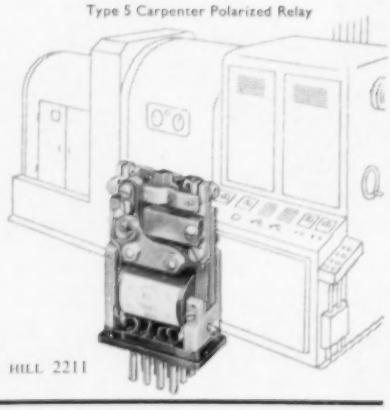

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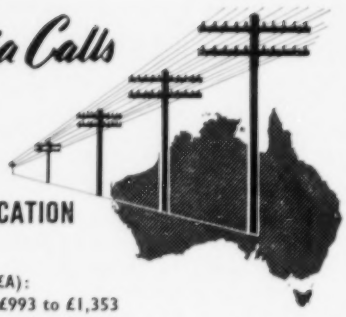

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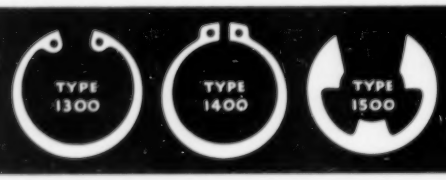

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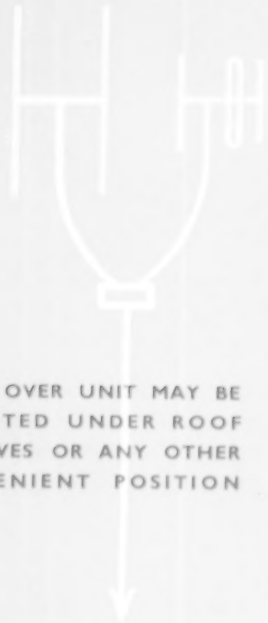

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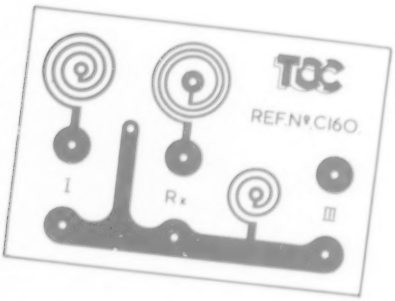

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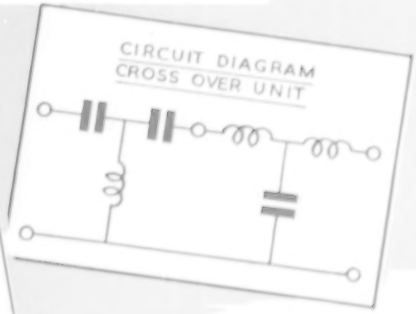

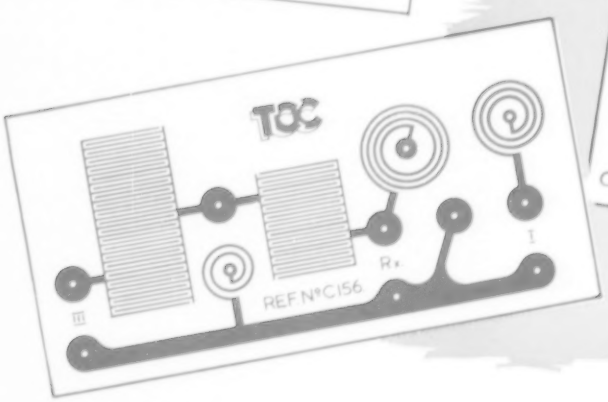

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